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(54) Title: CLOSED LOOP FEEDBACK SYSTEM FOR IMPROVED DOWN LINK PERFORMANCE

(57) Abstract: A method includes receiving at least two space-time coded signals from an antenna system associated with a first station, determining complex channel state information based on the received space-time coded signals, and sending the complex channel state information to the first station. In an alternative embodiment, a method includes transmitting at least two space-time coded signals in respective beams of a multi-beam antenna array, measuring a channel impulse response for each space-time coded signal at a second station, and sending an indicia of a selected set of least attenuated signals from the second station to the first station. The multi-beam antenna array is associated with a first station. The beams transmit a signature code embedded in each respective space-time coded signal, and the signature codes are orthogonal so that the second station can separate and measure the channel impulse response corresponding to each space-time coded signal. The space-time coded signals include the selected set of least attenuated signals and a remaining set of most attenuated signals. In an alternative embodiment, a method includes selecting at least two beams of plural beams formed by a multi-beam antenna array associated with a first station for transmission of a corresponding at least two space-time coded signals produced by a space-time encoder, determining a time delay associated with each of the at least two space-time coded signals as received in each respective beam, and setting into a variable delay line the time delay corresponding to each beam, each variable delay line being coupled between the multi-beam antenna array the space-time encoder.

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CLOSED LOOP FEEDBACK SYSTEM FOR IMPROVED DOWN LINK PERFORMANCE

BACKGROUND OF THE INVENTION

Field of the Invention

The present invention relates to a system to control down link signal transmission from a base station of a cellular radio system to a remote station. In particular, the invention relates to a closed loop phase and amplitude control system to adjust the phase and amplitude of down link transmitted signals.

Description Of Related Art

Cellular telephone systems are operated in environments that give rise to multi-path or reflections of the their signals, particularly in urban environments. In FIG. 1, base station transmitter 1 broadcasts its signal to remote station 2 (often mobile) along direct path 3. However, owing to the presence of tall building 4, transmitter 1 also broadcasts its signal to remote station 2 along indirect path 5, thus, giving rise to angular spread AS between the direction of arrival of direct path 3 at remote station 2 and the direction of arrival of indirect path 5 at remote station 2. Direct path 3 and indirect path 5 are recombined at remote station 2 where constructive and destructive superimposed signals cause random or what appears to be random fading and black out zones.

To reduce the effects of multi-path, known systems employ space time transmit diversity techniques. In FIG. 2, a known transmitter includes space time transmit diversity encoder 10, complex multipliers 12 and 14, and antennas 16 and 18. Space time

transmit diversity encoder 10 processes input signal S_{IN} into two channel signals CH_1 and CH_2. Multipliers 12 and 14 may impart a same orthogonalizing code OC on the two channel signals CH_1 and CH_2 to identify the two channels as containing information about input signal S_{IN}; however, different orthogonal identifiers (e.g., pilot sequences or training sequences) are applied to the different antenna signals so that the remote station can separately identify the signals from the two antennas. The multiplied channel signals are transmitted on respective antennas 16 and 18 substantially spaced apart by a distance (e.g., 20 wavelengths). Such spaced apart antennas are referred to as diversity antennas. In multi-path environments severe fading results when different propagation paths sum destructively at the receiving antenna. Using diversity antennas, the probability that both signals CH_1 and CH_2 will be in deep fade is low since the two signals are likely to propagate over different paths such as the multi-paths 3 and 5. Diversity antennas may be omni-directional antennas or antennas directed at antenna sectors with overlayed sectors. When diversity antennas are sufficiently separated in space, they can be regarded as orthogonal since they propagate signals in non-correlated channels (i.e., paths).

Input signal S_{IN} carries two symbols, S_1 and S_2, in time succession, the first symbol in symbol slot between 0 and T, and the second symbol in symbol slot between T and 2T. In FIG. 3, exemplary encoder 10 uses a QPSK modulation technique and includes time align register 20 and hold registers 22 to hold the two symbols. Base band carrier signal SBBC is inverted in inverter 24 to produce negative base band carrier -SBBC. QPSK modulator 26 encodes symbol S_1 onto base band carrier signal SBBC to produce a modulated first symbol, and QPSK modulator 28 encodes symbol S_1 onto negative base band carrier signal -SBBC to produce a modulated conjugate of the first symbol. QPSK modulator 30 encodes symbol S_2 onto base band carrier signal SBBC to produce a modulated second symbol, and QPSK modulator 32 encodes symbol S_2 onto negative base band carrier signal -SBBC to produce a modulated conjugate of the second symbol. The modulated conjugate of the second symbol is inverted in inverter 34 to produce a negative modulated conjugate of the second symbol. Analog multiplexer 36 switches the modulated first symbol into the first channel signal during the first symbol time slot (i.e., 0 to T, FIG. 2) and switches the negative modulated conjugate of the second symbol into the first channel signal during the second symbol time slot (i.e., T to

2T, FIG. 2) so that the signal on CH1 is $[S_1, -S_2^*]$. Analog multiplexer 38 switches the modulated second symbol into the second channel signal during the first symbol time slot (i.e., 0 to T, FIG. 2) and switches the modulated conjugate of the first symbol into the second channel signal during the second symbol time slot (i.e., T to 2T, FIG. 2) so that the signal on CH2 is $[S_2, S_1^*]$.

In FIG. 2, code OC consists of one code applied to both multipliers 12, 14 that is used as a CDMA spreading function to isolate the two signals transmitted from antennas 16 and 18 from other signals that may generate co-channel interference. Multipliers 12 and 14, multiply the first and second channel signals before being transmitted through antennas 16 and 18. RF up converters are not shown for simplicity.

At remote station 2, a receiver receives signals from both antennas 16 and 18 on a single antenna, down-converts the signals, despreads the signals using code OC, and recovers a composite of channels CH1 and CH2 as transmitted from antennas 16 and 18, respectively. In the first symbol time slot between 0 and T, the composite QPSK modulated signal R_1 is received (where $R_1 = k_{11}S_1 + k_{12}S_2$), and in the second symbol time slot between T and 2T, the composite QPSK modulated signal R_2 is received (where $R_2 = -k_{21}S_2^* + k_{22}S_1^*$ and the asterisk refers to a complex conjugate). Constant k_{11} is a transmission path constant from first antenna 16 to remote station 2 during the first time slot, constant k_{12} is a transmission path constant from second antenna 18 to remote station 2 during the first time slot, constant k_{21} is a transmission path constant from first antenna 16 to remote station 2 during the second time slot, and constant k_{22} is a transmission path constant from second antenna 18 to remote station 2 during the second time slot. The receiver derotates the channel to recover soft symbols S_1' and S_2', where

$$S_1' = k_{11}R_1 + k_{12}R_2 \text{ and } S_2' = k_{21}R_2^* + k_{22}R_1^*.$$

In this time space encoder technique, the first and second symbols are redundantly transmitted from separate antennas. The first symbol is encoded to be transmitted in both the first and second symbol time slots, and the second symbol is also encoded to be transmitted in both the first and second symbol time slots. The effect of this symbol recovery technique is that fading or drop out regions that may appear during one symbol time slot are less likely to appear during both symbol time slots when interleaving is also exploited. Interleaving is used before space-time coding to make adjacent bits less

correlated in time. Since the received symbols are recovered from received signals during both time slots, R_1 and R_2, the effect of fading is diminished.

However, the prior art does not exploit advantages provided by independent power and phase management of individual beams transmitted by different diversity type antennas to achieve greater spectral efficiency at the base station while minimizing co-channel interference. The prior art does not exploit advantages provided by spatial power management of independently directed beams to achieve greater spectral efficiency at the base station while minimizing co-channel interference.

SUMMARY OF THE INVENTION

It is an object to the present invention to improve the down link performance of a cellular radio system. It is another object to minimize undesired effects of fading and drop out.

These and other objects are achieved with a method that includes receiving at least two space-time coded signals from an antenna system associated with a first station, determining complex channel state information based on the received space-time coded signals, and sending the complex channel state information to the first station.

These and other objects are achieved with an alternative embodiment where the method includes transmitting at least two space-time coded signals in respective beams of a multi-beam antenna array, measuring a channel impulse response for each space-time coded signal at a second station, and sending an indicia of a selected set of least attenuated signals from the second station to the first station. The multi-beam antenna array is associated with a first station. The beams transmit a signature code embedded in each respective space-time coded signal, and the signature codes are orthogonal so that the second station can separate and measure the channel impulse response corresponding to each space-time coded signal. The space-time coded signals include the selected set of least attenuated signals and a remaining set of most attenuated signals.

These and other objects are achieved with an alternative embodiment where the method includes selecting at least two beams of plural beams formed by a multi-beam antenna array associated with a first station for transmission of a corresponding at least two space-time coded signals produced by a space-time encoder, determining a time

delay associated with each of the at least two space-time coded signals as received in each respective beam, and setting into a variable delay line the time delay corresponding to each beam, each variable delay line being coupled between the multi-beam antenna array and the space-time encoder.

BRIEF DESCRIPTION OF DRAWINGS

The invention will be described in detail in the following description of preferred embodiments with reference to the following figures wherein:

FIG. 1 is a schematic view of the radio environment in which the present invention is employed;

FIG. 2 is a block diagram of a known base station;

FIG. 3 is a block diagram of a known space time encoder;

FIG. 4 is a block diagram of a base station apparatus according to an embodiment of the present invention;

FIG. 5 is a block diagram of a base station apparatus according to another embodiment of the present invention;

FIG. 6 is a schematic diagram of a known hex corner reflector antenna system;

FIG. 7 is a schematic diagram of a known phase array antenna;

FIG. 8 is a schematic diagram in plan view of an exemplary three sector antenna system;

FIG. 9 is a schematic diagram of a known "Butler matrix" antenna;

FIG. 10 is a schematic diagram of a dual beam phase array antenna;

FIG. 11 is a block diagram of a base station apparatus according to another embodiment of the present invention;

FIG. 12 is a block diagram of a TDMA base station apparatus according to another embodiment of the present invention;

FIG. 13 is a block diagram of a closed loop beam power management system according to the present invention;

FIG. 14 is a block diagram of a radio system according the present invention;

FIGS. 15-17 are flow charts of methods of determining the angular power spectrum according to the present invention;

FIG. 18 is a graph of an angular power spectrum as received and/or computed by the present invention;

FIG. 19 is a block diagram of an embodiment of the present invention;

FIG. 20 is a flow chart of a method of feedback control according to the present invention;

FIG. 21 is a schematic view that illustrates the multi-path signal processed by the invention with a sector coverage antenna;

FIG. 22 is a graph showing the direct and multi-path signal of FIG. 21 that is received by a remote station;

FIG. 23 is a schematic view that illustrates the multi-path signal processed by the invention with a multi-beam antenna covering a sector;

FIG. 24 is a graph showing the direct signal and a delayed replica of the direct signal of FIG. 21 or 23 that is received by a remote station;

FIG. 25 is a graph showing the multi-path signal of FIG. 21 or 23 that is received by a remote station;

FIG. 26 is a block diagram of a base station apparatus with a programmable delay line according to an embodiment of the present invention;

FIG. 27 is a graph depicting a delay distribution profile according to the invention;

FIG. 28 is a flow chart of a set up method according to the present invention;

FIG. 29 is a flow chart of a time align method according to the present invention; and

FIG. 30 is a flow chart of a method of feedback according to the present invention.

DETAILED DESCRIPTION OF PREFERRED EMBODIMENTS

To achieve greater spectral efficiency of transmissions from the base station while minimizing co-channel interference, independent power management of individual beams transmitted by different antennas of the diversity antennas has been developed, and beamspace time encoder techniques has been developed to exploit angle of arrival

diversity and exploit spatial power management of independently directed beams. Beamspace time techniques differ from known space time encoder techniques by its use of two or more independently directed orthogonal beams to exploit power and beam width management and angle of arrival diversity. Orthogonal beams are separately identifiable to the receiver by using perpendicular polarization (two beam case), by using a different pilot code for each beam in a CDMA system in addition to the CDMA spread spectrum code that is common to all beams, by using a different spread spectrum code for each beam in a CDMA system without pilot codes, by using a different training sequence (e.g., pilot code) multiplexed into each beam in a TDMA system. Persons skilled in the art will appreciate that there are other orthogonal beam techniques not listed above or techniques that use different combinations of the above techniques that are equivalent for providing a means for the receiver at the remote station to separately identify the individual beams and recover the signals they carry.

Power management techniques to transmit different powers in different orthogonal beams improve spectral efficiency at the base station on a system wide basis by minimizing co-channel interference even when this power management control is applied to overlaid sector directed beams or omni directional beams of diversity antennas. However, with orthogonally coded beams that are directed differently, spatial power management of independently directed beams provides even further improvements. The relatively poor downlink performance of radio environments with large angular spreads is significantly improved by applying the beamspace time encoder techniques described herein.

In FIG. 4, a first embodiment of an improved transmitter 100 (referred to as power management of diversity antennas) includes known space time transmit diversity encoder 10 and complex multipliers 12 and 14. Improved transmitter 100 further includes scaling amplifiers 102 and 104 and diverse antennas 16 and 18. In a CDMA system, multipliers 12, 14 impart different spread spectrum codes to different beams so that a receiver at remote station 2 can discern the beams separately.

Although separate distinguishable spreading codes in a CDMA system are applied to multipliers 12, 14 as described here to create the orthogonal beams, it will be appreciated that any means to create orthogonal beams enable the separate power

management of the transmissions from the diversity antennas (i.e., overlaid coverage), or from controllable directional antennas for that matter. For example, in a CDMA system where the multipliers 12 and 14 are provided with the same spreading codes, another set of multipliers 12' and 14' (not shown) may be used for imparting pilot codes to the channel signals. Multipliers 12' and 14' are then provided with orthogonal pilot codes so the receiver in remote station 2 can separately discern the beams. In another variant, antennas 16 and 18 are constituted by a single antenna with two exciter elements arranged to generate two beams that are orthogonally polarized (e.g., polarized at a +/- 45 degree slant to the vertical or some other reference), but otherwise cover the same sector. Such beams are orthogonal, and transmissions over the respective signal paths experience uncorrelated fading.

Scaling control signals SA1 and SA2 separately control the amplification or attenuation achieved by separate scaling amplifiers 102 and 104, respectively. Scaling control signals SA1 and SA2 may be real to scale amplitudes, or imaginary to shift phases or complex with both real and imaginary components to both scale amplitudes and shift phases. It will be appreciated that the amplification may be applied at the output of encoder 10, before multipliers 12 and 14, after multipliers 12 and 14 or in antennas 16 and 18.

Antennas 16, 18 are diversity antennas that cover overlaid sectors or are omni-directional. This first embodiment differs from known space-time coded systems in that the power transmitted in each beam is separately controlled by SA1 and SA2.

In FIG. 5, a second embodiment of an improved transmitter 100 (referred to as angular spectral power management) includes known space time transmit diversity encoder 10 and complex multipliers 12 and 14. Improved transmitter 100 further includes scaling amplifiers 102 and 104 and controlled directional antennas 106 and 108. Unlike antennas 16 and 18 of FIG. 2, directional antennas 106 and 108 are directed toward direct path 3 and indirect path 5 (FIG. 1) or some other direction to cover angular spread AS or that portion of the angular power spectrum that exceeds a threshold as described herein. In a CDMA system, multipliers 12, 14 impart different spread spectrum codes to different beams or use other means so that a receiver at remote station 2 can discern the beams separately as described for the first embodiment using diversity

antennas. Scaling control signals SA1 and SA2 separately control the amplification or attenuation achieved by separate scaling amplifiers 102 and 104, respectively. Scaling control signals SA1 and SA2 may be real to scale amplitudes, or imaginary to shift phases or complex with both real and imaginary components to both scale amplitudes and shift phases. It will be appreciated that the amplification may be applied at the output of encoder 10, before multipliers 12 and 14, after multipliers 12 and 14 or in antennas 106 and 108. Although separate spreading codes in a CDMA system are applied to multipliers 12, 14 as described here to create the orthogonal beams, it will be appreciated that any means to create orthogonal beams enable the separate power management of the transmissions from the controlled directional antennas (i.e., directions selected as described herein).

In a third embodiment (referred to as directional diversity and not separately shown), amplifiers 102 and 104 of FIG. 5 are removed from transmitter 100 so that no differential amplification is achieved, and both channels CH1 and CH2 have balanced and equal amplification, but their signals are transmitted directionally through controlled directional antennas 106 and 108.

There are several means to implement controlled directional antennas. In FIG. 6, known hex controlled directional antenna system 6 includes six co-sited corner reflector antennas, such as corner reflector antenna 8, arranged in a circle and all depicted in plan view. Each corner reflector antenna 8 includes a single half wave dipole 12 as an exciter element and corner reflectors 14. Each corner reflector antenna 8 illuminates a 60 degree beam width in plan view. Hex diversity antenna system 6 has been shown to provide angle location information that gives the bearing angle from a base station to the remote station based on received signal strength at 820 MHz (Rhee, Sang-Bin, "Vehicle Location In Angular Sectors Based On Signal Strength", IEEE Trans. Veh. Technol., vol. VT-27, pp 244-258, Nov. 1978). Such co-sited corner reflector antennas could divide a 360 degree coverage into three sectors (120 degree antennas), four sectors (90 degree antennas), five sectors (72 degree antennas), eight sectors (45 degree antennas), or any convenient number of sectors that may be realizable.

In the second and third embodiments of the present invention, a controlled directional antenna system is used for cellular radio transmitter 1 (FIG. 1). A controlled

directional antenna system is defined as being capable of providing two or more distinguishable and separately controllable beams. It may be a single antenna with two or more exciter elements arranged to generate two or more beams (e.g., arranged to generate two discernable beams respectively polarized at a +/- 45 degree slant to the vertical, but otherwise cover the same sector). It may be a multi-antenna system to generate beams that cover different sectors. For example, the controlled directional antenna system may advantageously be a hex corner reflector system, such as the antenna system depicted in FIG. 6. The controlled directional antenna system is used in a receive mode to determine the angle location of remote station 2 based on a signal transmitted from remote station 2. The two sectors with the strongest received signals are identified as the likely direction of arrival of direct path 3 and indirect path 5 (see FIG. 1). The antennas illuminating these two sectors are selected to be directional antennas 106 and 108 of the second and third embodiments of the present invention (FIGS. 4 and 5). Alternatively, the respective directions of arrival may be determined based on a calculation of the angular power spectrum as discussed below.

In FIG. 7, known stearable beam phased array antenna 20 includes an array of exciter elements 22 (e.g., half wave dipole) disposed to be spaced from ground plane or reflector plane 24. FIG. 7 depicts eight radiating elements, but more or fewer elements may be used. Each exciter element 22 is fed with a signal from a corresponding phase shifter 26. Each phase shifter 26 alters the phase and attenuates (or amplifies) the amplitude of signal S according to a corresponding individual control portion of control signal C. For example, control signal C includes 8 phase shift parameters and 8 attenuation parameters. Each phase and amplitude parameter individually controls the phase and amplitude radiated from a corresponding element of the eight exciter elements of antenna 20. The angular beam width of such an antenna is limited by the ratio of the wavelength of the signal being radiated divided by the aperture dimension D; however, by controlling signal amplitudes on exciter elements 22 as distributed across the antenna with what is called a weighting function, the beam may be shaped to broaden the beam, flatten the center of the beam and/or suppress side lobes. By controlling the gradient of the phase at the exciter elements across the antenna, the beam may be electronically directed to point in a controlled direction.

In a variant of the second and third embodiments, the antenna system for transmitter 1 (FIG. 1) includes plural phased array antennas 20 organized in a multi-antenna system. In FIG. 8, an exemplary multi-antenna system may include three antennas (taken to be phased array antennas 20) arranged to point outward in equally spaced angular direction so that the three phased array antennas 20 are formed into the antenna system at the base station. Each antenna 20 is designed to cover a 120 degree sector. The base station locates the remote station by electronically scanning antenna 20. Amplitude weights for each radiating element are preferably set to a maximum and are all equal so that the antenna provides its narrowest beam (most directional beam). The receive beam is scanned in steps by first computing the phase parameters for control signal C that represent a gradient in phase across the antenna to achieve a desired beam point, and then controlling antenna 20 to point in the desired direction. Second, a receiver at transmitter 1 (FIG. 1) detects any received signal strength. The steps of pointing a receive beam and detecting a signal strength are repeated at each of several beam positions until the entire sector covered by antenna 20 has been scanned. In this way, the angle location of remote station 2 is determined to a precision limited only by the narrowest achievable beam width of antenna 20. Once the location of direct path 3 and indirect path 5 are determined to be in different sectors (e.g., 120 degree sectors), antennas 106 and 108 (FIG. 5) are selected from the plural antennas 20 of the antenna system that are closest to direct path 3 and indirect path 5, and within the sector covered by each selected antenna 20, the phase gradients that define beams pointing at the angle locations for direct path 3 and for indirect path 5 are determined. Alternatively, when paths 3 and 5 lie in a single sector, two transmitting beams can be formed within the single sector to be directed along paths 3 and 5 if the antenna system is capable of forming the two beams in the single sector (see discussion below with respect to FIG. 10).

In FIG. 9, antenna system 30 includes four radiating elements 32 disposed to be spaced from ground plane or reflector plane 34. Each radiation or exciter element 32 is fed with a signal from known Butler matrix 36. The Butler matrix provides phase shifting and combination functions that operate on signals S1, S2, S3 and S4 so that the radiation from the four exciter elements 32 combine to generate four fixed angularly

directed and orthogonal beams B1, B2, B3 and B4. In general, a Butler matrix performs a Fourier processing function to feed M radiating elements so as to form M fixed and orthogonal beams ("angular bins"). For example, in antenna system 30, signal S1 is transmitted only in first beam B1, signal S2 is transmitted only in second beam B2, signal S3 is transmitted only in third beam B4, and signal S4 is transmitted only in fourth beam B4. A switching matrix may be used to direct desired signals (e.g., the signals CH1 and CH2 of FIG. 5) onto any of the lines for signals S1, S2, S3, and S4 and from there into respective beams B1, B2, B3 and B4.

In a variant of the second and third embodiments, the antenna system for transmitter 1 (FIG. 1) includes plural "Butler matrix" antennas 30 organized in a multi-antenna system. In FIG. 8, an exemplary multi-antenna system includes three antennas (taken here to be "Butler matrix" antennas 30) arranged to point outward in equally spaced angular direction so that the three "Butler matrix" antennas 30 are formed into the antenna system at the base station. Each antenna 30 is designed to cover a 120 degree sector with, for example, four beams. The base station locates the remote station by electronically switching between the four beams (each 30 degrees) of each of the three antennas 30 and detecting the signal strength received. In this way, the angle location of remote station 2 is determined to a precision of one beam width of antenna 30. Once the locations of direct path 3 and indirect path 5 are determined, antennas 106 and 108 (FIG. 5) are selected from the two different "Butler matrix" antennas 30 that make up the antenna system for transmitter 1 (FIG. 1) if direct path 3 and indirect path 5 lie in different sectors. The two particular "Butler matrix" antennas 30 are selected to cover the sectors that are closest to direct path 3 and indirect path 5, and from there, a particular beam within each selected antenna 30 is selected that most closely aligns with the path. Alternatively, antennas 106 and 108 may be selected to be different beams of the same "Butler matrix" antenna 30. Within the sector covered by each antenna 30, the beam pointing at the angle location for each of direct path 3 and indirect path 5 is selected by a switch matrix (not shown).

In FIG. 10, antenna 40 is a modified version of phased array antenna 20 to provide two independently steerable and shapable beams. Antenna 40 includes an array of exciter elements 42 (e.g., half wave dipole) disposed to be spaced from ground plane or reflector

plane 44. FIG. 10 depicts eight radiating elements, but more or fewer elements may be used. However, unlike antenna 20, each exciter element in antenna 40 is fed by a signal from a corresponding summer 48. Each summer 48 superimposes (e.g., adds) signals from two corresponding phase shifters 46-1 and 46-2. All phase shifters 46-1 form a first bank of phase shifters, and all phase shifters 46-2 form a second bank of phase shifters. Each phase shifter 46-1 in the first bank alters the phase and attenuates (or amplifies) the amplitude of signal S1 according to a corresponding individual control portion of control signal C1. For example, control signal C1 includes 8 phase shift parameters and 8 attenuation parameters to individually control the phase and amplitude output from the corresponding phase shifter 46-1. Correspondingly, each phase shifter 46-2 in the second bank alters the phase and attenuates (or amplifies) the amplitude of signal S2 according to a corresponding individual control portion of control signal C2. For example, control signal C2 includes 8 phase shift parameters and 8 attenuation parameters to individually control the phase and amplitude output from the corresponding phase shifter 46-2. Summers 48 combine the outputs of respective phase shifters 46-1 and 46-2 and provide the combined signal to radiating elements 42. In this way, control signal C1 controls a first beam that radiates signal S1, and control signal C2 simultaneously controls a second beam that radiates signal S2.

In a variant of the second and third embodiments, the antenna system for transmitter 1 (FIG. 1) includes plural phased array antennas 40 organized in a multi-antenna system. In FIG. 8, an exemplary multi-antenna system includes three antennas (taken here to be phased array antennas 40) arranged to point outward in equally spaced angular direction so that the three phased array antennas 40 are formed into the antenna system at the base station. Each antenna 40 is designed to cover a 120 degree sector with two independently shapable and steerable beams. The base station locates the remote station by electronically scanning a beam of antenna 40 as discussed above with respect to antenna 20 (FIG. 7). Once the location of direct path 3 and indirect path 5 are determined, antennas 106 and 108 (FIG. 5) are selected from the plural antennas 40 of the antenna system that are closest to direct path 3 and indirect path 5, and within the sector covered by each selected antenna 40, the phase gradients that define beams pointing at the angle location for direct path 3 and for indirect path 5 are determined.

Alternatively, antennas 106 and 108 may be selected to be different beams of the same dual beam antenna 40. In FIG. 11, antennas 106 and 108 (FIG. 5) are implemented in separate beams (i.e., beams 1 and 2) of dual beam antenna 40, and scaling amplifiers 102 and 104 (of FIG. 5) are not needed since the scaling function may be achieved by scaling the amplitude coefficients of control signals C1 and C2 (FIG. 10).

In a fourth embodiment, the base station uses a time division multiple access (TDMA) transmitter instead of a spread spectrum CDMA transmitter. In FIG. 12, training sequence TS1 is modulated in QPSK modulator 101 and from there fed to a first input of multiplexer 105, and training sequence TS2 is modulated in QPSK modulator 103 and from there fed to a first input of multiplexer 107. Training sequences TS1 and TS2 are orthogonal and provide the means by which remote station 2 can discern between the beams in much the same was as pilot codes help distinguish beams in a CDMA system. In the TDMA system, multipliers 12 and 14 (of FIGS. 4, 5 and 11) are omitted and channel signals CH1 and CH2 are fed to second inputs to multiplexers 105 and 107, respectively. In this fourth embodiment amplifiers 102 and 104 independently amplify or attenuate the outputs of respective multiplexers 105 and 107. The outputs of amplifiers 102 and 104 are fed to the antenna system (through up converters, etc., not shown). The antenna system may provide the overlaid coverage of diversity antennas 16, 18 (FIG. 4) as in the first embodiment or may provide controlled directional coverage of directional antennas 106, 108 (FIGS. 5 and 11) as in the second and third embodiments. Moreover, in the case of controlled directional coverage, a variant may be to forego power management and omit amplifiers 102, 104 and rely on angle (beam) diversity by steering beams from directional antennas 106, 108. A data slot in a time division system may include, for example, 58 data bits followed by 26 bits of a training sequence followed by 58 data bits as in a GSM system. The training sequence identifies the source of signal S_{IN} and the individual beam to remote station 2 so that the remote station can separately discern the beams. In this way, remote station 2 can separately receive the two beams using the training sequences, instead of using orthogonal spreading codes OC as in a CDMA system.

Although two beams are discussed, extensions to higher order coding techniques with more beams are straightforward. For example, four symbols (S_1, S_2, S_3, S_4) encoded

into four channel signals (CH1, CH2, CH3, CH4) in four symbol time slots so that the original symbols are recoverable from the encoded channel signals. The four channel signals are then transmitted from the base station in four beams, each beam corresponding to a channel signal of the channel signals CH1, CH2, CH3, and CH4. Although QPSK modulation techniques are discussed herein, extensions to other PSK modulation techniques are straightforward, and extensions to other modulation techniques (e.g., QAM) are equally useable.

In FIG. 13, a closed loop control system to manage transmit powers is depicted as process S10. In step S102, the base station selects the power level to be transmitted from each antenna. For example, in a two antenna system, the base station selects powers P1 and P2 based on the total power (i.e., P1 + P2) as defined by a conventional power control loop (e.g., a control loop typical to a CDMA system) and the relative powers (i.e., P1/P2) as defined by power control coefficients measured at remote station 2. In step S104, a value representing the selected transmit power level is sent to the remote station in a signaling channel. In step S106, the power level received at the remote station from each antenna radiation pattern is measured, and corresponding power control coefficients are determined. The power control coefficients for each antenna radiation pattern are determined at remote station 2 to be proportional to the received power at remote station 2 divided by the transmitted power as indicated by the power level value that is sent to the remote station in a signaling channel. In step 106 the power control coefficients are sent from the remote station to the base station in a signaling channel. In step S108, the power control coefficients from step S106 are compared for each antenna. In step S110, adjustments in transmit signal power are determined according to the comparision of step S108. The adjustments are made to increase transmit powers sent in channels that have favorable transmission qualities and reduce transmit powers in channels that have poor transmission qualities. Then, in step S102 at the begining of the cycle, the base station selects adjusted transmit powers to form the basis for the powers to be transmitted from the antennas during the next cycle of the closed loop beam power management. The loop cycle delay may be one time slot as in a third generation TDMA system.

Alternatively, the remote station may compare (in step S108) the power control coefficients for each antenna from step S106 and then compute power coefficient

indicator information to be sent from the remote station to the base station in an up link signaling channel. For example, a ratio of the power control coefficients (e.g., P1/P2 in a two antenna case) may be advantageously computed as the power coefficient indicator information and transmitted in the up link direction. Or the power coefficient indicator information may be quantized value of the ratio (e.g., a single bit indicating whether P1 > P2 or not).

Alternatively, in step S104, the selected transmit power is saved for a cycle time of the closed loop control system. For example, in a two antenna system, the base station selects powers P1 and P2 based on the total power (i.e., P1 + P2) as defined by a conventional power control loop (e.g., a control loop typical to a CDMA system) and the relative powers (i.e., P1/P2) as defined by power control coefficients measured at remote station 2. In step S106, the power levels received at the remote station from each antenna radiation pattern are measured at remote station 2 and sent as power control coefficients in an up link signaling channel from remote station 2 to base station 1. The power control coefficients are normalized to their respective transmit powers as saved in step S104. In step S108, the normalized power control coefficients from step S106 are compared at the base station for each antenna. In step S110, adjustments in transmit signal power are determined according to the comparison of step S108. Then, in step S102 at the beginning of the cycle, the base station selects adjusted transmit powers to form the basis for the powers to be transmitted from the antennas during the next cycle of the closed loop beam power management.

In FIG. 14, a cellular radio system with closed loop beam power management controls includes base station 210 and remote station 230. Base station 210 includes space-time encoders 212 to encode a stream of symbols into first and second space-time coded signals, antenna system 216, transmitter 214 to transmit the first and second space-time coded signals at respective first and second initial transmit powers from the antenna system so as to form respective first and second radiation patterns, base station receiver 220 to receive power coefficient indicator information from the remote station, and power management controller 222 to determine first and second adjusted transmit powers based on the respective first and second initial transmit powers and the power coefficient indicator information.

Antenna system 216 may include plural antennas where each antenna is an antenna that generates either a substantially omni-directional radiation pattern or a radiation pattern directed to a sector. Omni-directional antennas are advantageously spaced apart. Antenna system 216 may form the first and second radiation patterns as orthogonal radiation patterns capable of being separately received at the remote station. Alternatively, transmitter 214 includes a circuit to process the first and second space-time coded signals so that the signals transmitted from the antenna system are orthogonal and can be separately received at the remote station.

Antenna system 216 is capable of generating plural beams (i.e., a multi-beam antenna) and the base station include antenna control 218 to control the multi-beam antenna to form the plural beams. In one embodiment, the multi-beam antenna may be a multi-port Butler matrix antenna, and in this case, transmitter 214 will include amplifiers to scale the first and second space-time coded signals to form respective first and second scaled space-time coded signals based on the respective first and second adjusted transmit powers, and antenna control 218 will include a switch to couple the first and second scaled space-time coded signals into respective first and second input ports of the Butler matrix antenna to form the respective first and second beams.

Alternatively, the multi-beam antenna includes a phased array antenna system, and antenna control 218 includes a beam steering controller to form first and second weighting functions. The beam steering controller includes logic to input the first and second weighting functions into the phased array antenna system to scale antenna gains of the respective first and second beams based on the respective first and second adjusted transmit powers without scaling amplifiers in transmitter 214. The phased array antenna system may include either a plural beam phased array antenna (e.g., 40 of FIG. 10) or plurality of phased array antennas (e.g., 20 of FIG. 7).

In some embodiments, the power coefficient indicator information includes first and second power control coefficients, and base station receiver 220 receives up link signaling information and detects values of the first and second power control coefficients in the up link signaling information.

Power management controller 222 includes a circuit (e.g., logic or a processor) to determine the first adjusted transmit power to be greater than the second adjusted

transmit power when the indicated first path attenuation characteristic (or first power control coefficient) is less than the indicated second path attenuation characteristic (or second power control coefficient).

Remote station 230 includes remote station receiver 234, detector 236, power measurement circuit 238 and processor 240. Receiver 234, detector 236, power measurement circuit 238 and processor 240 constitute a circuit by which remote station 230 can determine an indicated path attenuation characteristic based on a power received from the first radiation pattern and measured in circuit 238 and an initial transmit power determined in detector 236. With this circuit, remote station 230 can determine an indicated first path attenuation characteristic for a first radiation pattern of antenna system 216 and an indicated second path attenuation characteristic for a second radiation pattern of system 216 since the two radiation patterns are separately receivable. Detector 236 determines the initial transmit power, power measurement circuit 238 measures the power received from the radiation pattern as received by receiver 234, and processor 240 determines a power control coefficient to be proportional to the power received divided by the value of the initial transmit power. Power measurement circuit 238 measures an instantaneous power received, or in an alternative embodiment, measures an averaged power received, or in an alternative embodiment measures both and forms a combination of the instantaneous power received and the average power received. Remote station 230 further includes transmitter 242 to send values the power coefficient indicator information or of the indicated first and second path attenuation characteristics to the base station.

In a variant, processor 240 forms the power coefficient indicator information as a ratio of the indicated first path attenuation characteristic divided by the indicated second path attenuation characteristic. In an alternative variant, processor 240 forms the power coefficient indicator information with a first value when the indicated first path attenuation characteristic is less than the indicated second path attenuation characteristic and to form the power coefficient indicator information with a second value when the indicated first path attenuation characteristic is greater than the indicated second path attenuation characteristic.

In an exemplary embodiment, the base station transmits a first signal at first predetermined signal power P1 from the first antenna, and a receiver in remote station 2 determines first power control coefficient PCC1 to be a power received from the first antenna at the remote station. The base station also transmits a second signal at second predetermined signal power P2 from the second antenna, and a receiver in remote station 2 determines second power control coefficient PCC2 to be a power received from the second antenna at the remote station.

Both the first and second signals are transmitted simultaneously from respective first and second antennas in ordinary operation at their respective predetermined power levels. The transmit powers are distinguishable at remote station 2 by use of different orthogonal codes OC in multipliers 12 and 14 (FIGS. 4, 5 and 11) or by use of orthogonal training sequences as may be used in a TDMA base station (FIG. 12). The receiver in remote station 2 determines the signal power received from each antenna and transmits a value representing these received signal powers to the base station in a portion of the up link signaling data as separate power control coefficients PCC1 and PCC2 or as a relative power control coefficient PCC1/PCC2.

In a preferred embodiment, the base station first transmits signals in ordinary operation from the plural antennas at selected powers that may be unequal (S102). In one variant, the base station sends the power levels selected to be transmitted from each of the plural antennas in a down link signaling channel. The remote station (1) receives the base station's selected power levels (S104), (2) determines the signal powers received from the antennas (S106), and (3) compares the power transmitted from the base station from each antenna to the powers received at the remote station to determine the relative attenuations in the down link paths (S108) as the ratio of the received power to the corresponding transmitted power. The remote station sends this ratio determined for each antenna as power control coefficients back to the base station in the up link signaling data. Then, the base station adjusts the power allowed to be transmitted from the base station from each antenna according to the determined relative attenuations for all further down link transmissions (S110).

In another variant, (1) the remote station determines power control coefficients to be the signal powers received from the antennas (S106), and (2) the remote station

sends the power control coefficients back to the base station in the up link signaling data. Then, the base station (1) adjusts for closed loop time delays in its receipt of the power control coefficients from remote station 2 (S104), (2) compares the power transmitted from the base station from each antenna to the power control coefficients received at the remote station to determine the relative attenuations in the down link paths (S108), and (3) adjusts the power allowed to be transmitted from the base station from each antenna according to the determined relative attenuations for all further down link transmissions (S110).

In either variant, the power allowed to be transmitted from an antenna will be greater for antennas associated with paths determined to posses a lesser path attenuation. For example, an indicated path attenuation characteristic is advantageously determined to be the ratio of the power received at remote station 2 to the power transmitted from base station 1. In this way, little or no power is transmitted in a path that is not well received by remote station 2, while a greater power is transmitted in a path that is well received by remote station 2. In many multi-path environments, increasing power transmitted in a path that has too much attenuation does little to improve reception at remote station 2, but such increased power would contribute to co-channel interference experienced by other remote stations. To improve the overall cellular radio system, the paths with the least attenuation are permitted the greatest transmit beam powers. The base station adjusts the power transmitted from each antenna by control scaling signals SA1 and SA2 (FIGS. 4 and 5) or by controlling the overall antenna gain for each beam by adjusting the amplitude parameters in control signal C (of FIG. 6) or in signals C1 and C2 (of FIG. 9).

In an embodiment of this closed loop method of power control, the remote station determines which antenna (or beam) is associated with the least attenuation path. The remote station sends an indication of which antenna (or beam) is favored (i.e., least attenuation) back to the base station in an up link signaling path. To conserve the number of bits sent in this up link signaling path, the remote station preferably determines the favored antenna and indicates this by a single bit (i.e., a "0" means antenna 16 is favored and a "1" means antenna 18 is favored, see FIG. 4). The base station receives this single bit indicator and applies it to determine a predetermined relative power balance. For

example, it has been determined that applying 80% of full power to antenna 16 (e.g., when this is the favored antenna) and 20% of full power to antenna 18 consistently provides better performance than applying 100% of full power to antenna 16 and no power to antenna 18. Thus, the base station receives the single bit relative power indicator and selects the relative power P1/P2 for antennas 16 and 18 to be 80%/20% for a "1" indicator bit and 20%/80% for a "0" indicator bit.

In slowly varying radio environments, the coefficients (or any related channel information) can be parsed into segments, and the segments (containing fewer bits than the entire coefficient) can be sent to the base station in the up link signaling data using more up link time slots. Within a segment (perhaps plural TDMA time slots), the most significant bits are preferably transferred first, and these course values are gradually updated to be more precise using consecutive bits. Conversely, in rapidly varying radio environments, a special reserved signaling symbol may indicate the use of one or more alternative compressed formats for the up link transmission of the coefficients where an average exponent of all of the coefficients is transmitted (or presumed according to the signaling symbol) in the up link, and then only the most significant bits of the coefficients are then transmitted (i.e., truncating the less significant bits). In the extreme, only one bit is transmitted in the up link direction indicating that the power control coefficient is 1 (e.g., 80% of full power transmission) when the down link channel is good, and indicating that the power control coefficient is 0 (e.g., only 20% of full power transmission) when the associated channel is not adequate.

This closed loop control over beam power management is self adapting. If power control coefficients are up linked to the base station that cause over compensation in beam power, this closed loop control system will correct for this during the next closed loop control cycle. Persons skilled in the art will appreciate that other data compression techniques may be employed in the up link signaling to adjust to rapidly varying radio environments. Similarly, persons skilled in the art will appreciate that the remote station, not the base station, may compute commands to the base station to increase, or decrease, the power in specific beams.

In an alternative variant suitable for slowly varying radio environments, the first and second beams may be sequentially transmitted at their respective predetermined

power levels in a calibration mode. In such a variant, only one beam is transmitted at a time so that the remote station need not employ orthogonal codes OC or orthogonal pilot signals to determine from which beam the received signal strength (e.g., power control coefficient) has been received. Once the channel attenuation is determined, signal S_{IN} is sent using the beamspace time coding technique.

In addition to embodiments that rely on amplifiers 102 and 104 or beam gain in phase array antennas to control closed loop power management, another embodiment relies on angular diversity management and/or beam width management with the power management being omitted. Yet another embodiment relies on both the power management and either angular diversity management, beam width management, or both.

The performance of beamspace time coding techniques depends at least in part on angular spread AS that characterizes the radio environment and how the base station adapts the beams to match the angular spread. Down link performance is generally improved when the down link beams are directed at angles of arrival at which sharp peaks occur in an angular power spectrum of a signal from a remote station. The sharp peaks suggest good transmission along the indicated path (e.g., likely direction of paths 3 and 5). However, sharp peaks may not always be found. When the angular power spectrum is diffuse and sharp peaks cannot be found, an estimate of angular spread AS is made, and the plural beams used for down link transmissions are allocated to approximately cover the angular spread. In this way the down link transmission spatially matches the total channel as determined by the angular spread.

The circuit to measure the angular power spectrum includes receiver 220 (FIG. 14) and such signal and data processing circuitry as is required to determine the angular power spectrum and peaks therein as discussed below. When a peak in the angular power spectrum is detected, an angular position is defined by the peak. Then, to direct the beam direction toward an angular position as detected, antenna controller 218 computer an array steering vector to input into antenna system 216 (FIG. 14). When an excessive number of peaks are detected in the angular power spectrum, power management controller 222 (FIG. 14) selects the angular directions to be used to form beams. Power management controller 222 may select beams directions toward specific angle of arrival paths (i.e., peaks), or power management controller 222 may select beam directions, and

possibly beam widths, so as to cover a detected angular spread. The selected directions are provided to antenna controller 218 to form the beam commands to the antenna system.

In systems using frequency division duplexing, the up link and down link transmissions take place at different frequencies. There is no guarantee that peaks measured in the up link power spectrum will occur at angle that correspond to angles with good transmission performance in the down link direction. However, by employing either angle diversity management or beam width management or both, there will be a greater likelihood of producing a good down link transmission.

Both angular diversity and beam width management require a measurement of the angular power spectrum in one form or another. The remote station broadcasts an up link signal in its normal operation (e.g., signaling operation), the antenna system at the base station receives the signal, and the base station determines an angular power spectrum (i.e., a received power as a function of bearing angle in a plan view). FIG. 18 is a graph depicting the angular location of signal power received from remote station 2. In FIG. 18, discrete power measurements at each of 12 angular locations are shown based on, for example, twelve fixed location antenna beams pointed at 30 degree intervals in the antenna system for base station 1. The exemplary 12 beam antenna system may include three Butler matrix antennas, triangularly arranged, to form the 12 beam antenna system where each Butler matrix antenna forms four beams. While a 12 beam antenna system is considered in this example, it will be appreciated that any number of beams in an antenna system may be applied to the present invention (e.g., 24 beams, etc).

Alternatively, the antenna system may include three phased array antennas, triangularly arranged, to form an antenna system capable of forming the 12 beam where each phased array antenna forms a steerable beam with a beam width of 30 degrees so as to permit scanning over four beam positions. The 12 beam antenna system may also include 12 antennas of any type that have a 30 degree beam width and are angularly disposed at 30 degree increments around a 360 degree sector. While a 12 beam antenna system is considered in this example, it will be appreciated that any number of beams in an antenna system may be applied to the present invention (e.g., 24 beams, etc).

An antenna system based on a phased array antenna provides an opportunity to generate a more interpolated angular power spectrum (e.g., G1 of FIG. 18) by steering the antenna beam to point at as many angular positions as desired to generate the angular power spectrum. Power management controller 222 (FIG. 14) generates the angular power spectrum in process S20 (FIG. 15) by looping on θ in steps S20A and S20B and determining the angular power in step S21. Given the angle θ, power management controller causes antenna controller 218 (FIG. 14) to computer an array steering vector and point the antenna (step S211 of FIG. 16). The phased array antenna then receives a signal in receiver 220 (FIG. 14) from remote station 2 in each radiating element of the phased array antenna to form a signal vector in step S212 of FIG. 16. Each radiating element is preferably spaced apart from an adjacent element by one-half of the wavelength. For example, if a phased array antenna were to include 12 radiating elements (only 8 radiating elements are shown in antenna 20 of FIG. 7), the signal received in each of the 12 radiating elements would be sampled to form a measured signal vector. The sampled signal is preferably a complex value having amplitude and phase information. The signals from each of the 12 radiating elements are formed into a 12 element received signal vector as column vector $\hat{x}$. Next, the complex conjugate transpose of received signal vector $\hat{x}$ is formed as row vector $\hat{x}^H$, and the spatial covariance matrix of the received signal, $R=\hat{x}\hat{x}^H$, is calculated in step S213 (FIG. 16). When received signal vector $\hat{x}$ is 12 elements long, then the spatial covariance matrix of the received signal, $R=\hat{x}\hat{x}^H$, will be a 12 by 12 matrix.

Array steering vector $\bar{a}(\theta)$ is a column vector with one vector element for each radiating element of the phased array antenna. For example, if the phased array antenna were to include 12 radiating elements (e.g., half dipoles), array steering vector $\bar{a}(\theta)$ would include 12 vector elements. Array steering vector $\bar{a}(\theta)$ is constant C of

FIG. 7, and it is used to point the beam of the phased array antenna toward bearing angle θ. Each vector element is given by:

$$\bar{a}_m(\theta)=\exp(-j\times k\times m\times d\times \sin\theta) \; ,$$

where k is 2π divided by the wavelength, m is an index from 0 to M (e.g., from 0 to 11 for a 12 element antenna) defining a number associated with the radiating element of the phased array antenna, d is the separation between radiating elements of the phased array antenna (preferably one-half of the wavelength) and θ is the bearing angle of the antenna beam formed.

Each vector element of array steering vector $\bar{a}(\theta)$ is a corresponding vector element of constant C as depicted in FIG. 7 so that the full vector combines to define an angle of arrival θ of the received signal in the receive beam, where θ is an angle with respect to a convenient reference direction of the phased array antenna. The complex conjugate transpose of array steering vector $\bar{a}(\theta)$ is row vector $\bar{a}(\theta)^H$.

The product, $\hat{x}\hat{x}^H\bar{a}(\theta)$, is still a column vector with one vector element for each radiating element of the phased array antenna. The product, $\bar{a}(\theta)^H\hat{x}\hat{x}^H\bar{a}(\theta)$, is a single point, a scalar, determined at step S214 (FIG. 16) to give the value of the angular power spectrum P(θ) at the angle of arrival θ. Thus, the angular power spectrum P(θ) is depicted in FIG. 18 at G1 and is computed to be:

$$P(\theta)=\bar{a}(\theta)^H\hat{x}\hat{x}^H\bar{a}(\theta)$$

where $\bar{a}(\theta)$ is an array steering vector, $\hat{x}$ is the received signal vector, $\hat{x}\hat{x}^H$ is the spatial covariance matrix of the received signal, and H denotes the complex conjugate transpose.

The above described equation for computing the array steering vector assumes the half wavelength spaced radiating elements are arrayed linearly. However, it will be

appreciated by persons skilled in the art how to compute an array steering vector for radiating elements arrayed long a curved path. Three slightly "bowed out" antenna arrays may advantageously be employed in the antenna system depicted in FIG. 8. In fact, the antenna arrays may be severely "bowed out" so as to form a circle (e.g., FIG. 6). It will be appreciated by persons skilled in the art that computation of an array steering vector for such severely curved arrays of radiating elements will advantageously employ amplitude control as well as phase control in the array steering vector.

To provide improved performance the angular power spectrum is determined by averaging repeated measurements. In FIG. 17, the array steering vector is prepared and the antenna beam is pointed in step S211. The plural measurements are made by looping in steps S215A and S215B. Within this loop, received signal vector $\hat{x}$ is repeatedly measured in step S216 and the covariance matrix R is repeatedly determined and saved in step S217. Then, an average covariance matrix is determined in step S218, and angular power spectrum P(θ) is determined in step S214. This averaging determination is repeated several times over a time interval for each predetermined direction θ. In this way, fast fading phenomena are averaged out. The time period must be short enough that a mobile remote station 2 will not change position sufficiently to change the beam in which it is located during the averaging period. This time period, is preferably larger than the channel coherence time to average out fast fading effects. While the channel coherence time is not rigorously and universally defined, it may be taken to be proportional to and approximately equal to an inverse of the Doppler spread.

The Doppler spread is more rigorously defined. Due to a relative velocity between the base station and a mobile remote station, there will be a physical shift in the received frequency with respect to the transmitted frequency. The Doppler spread is twice this frequency shift. For example, the Doppler frequency shift is the ratio of the relative velocity to the wavelength (in like units, meters/second divided by meters or feet/second divided by feet, etc.). If a mobile remote station is traveling 13.9 meters/second (about 50 km/h) and the wavelength is about 0.15 meters (e.g., 2,000 MHz signal with the speed of light equal to 300,000,000 meters per second), then the Doppler frequency shift is 92.7 Hz, the Doppler spread is 185 Hz, and the channel coherence time

is about 5.4 milliseconds. It can be easily verified that at a relative velocity of 40 meters per second (about 144 km/h) the channel coherence time is about 1.9 milliseconds, and that at a relative velocity of 1 meter per second (about 3.6 km/h) the channel coherence time is about 75 milliseconds.

The averaging time interval is preferably set to be greater than an inverse of the Doppler spread and less than a time in which a mobile station moving at an expected angular speed moves one-half of a beam width of the base station antenna system. The base station knows the remote station's range or can infer the range from signal strength. The base station is designed to communicate with mobile stations that can move at speeds up to a predetermined speed. This speed divided by the range may be taken to be the angular speed if the mobile station is moving radially around the base station. Setting the averaging interval to be a half beam width divided by the angular speed provides an estimate of the time in which a mobile remote station 2 will not change position sufficiently to change the beam in which it is located during the averaging period.

The time period over which the power P(θ) is averaged is usually much greater than the channel coherence time. For example, in a wide band CDMA system operating in an environment with a high incidence of multi-path reflections (e.g., urban environment), the average period could be tens of time slots. For indoor environments with a high incidence of multi-path reflections, the mobile is much slower and the averaging period can be much longer.

The base station computes the angular power spectrum and determines whether or not sharp peaks are indicated in the power spectrum. When sharp peaks are indicated, the angle location of each peak is determined. When the power spectrum is diffuse and no sharp peaks are indicated, the base station determines angular spread AS by first determining the angles at which the received angular power spectrum exceeds a predetermined threshold (G2 in FIG. 18). The threshold may also be adaptable based on the radio environment (e.g., signal density) detected by base station 1.

Sharp peaks in the angular power spectrum may be detected by, for example, using a two threshold test. For example, determine a first continuous angular extent (in degrees or radians) at which the power spectrum exceeds a first threshold G3. Then, determine a second continuous angular extent at which the power spectrum exceeds a

second threshold G2 (lower than first threshold G3). When the ratio of the first angular extent divided by the second angular extent is less than a predetermined value, peaks are indicated.

When peaks are indicated, angle diversity management (i.e., the management of the direction of arrival of the beams) is invoked, and possibly beam width management is invoked. The sharpness of the spectral peaks may be determined by comparing the angular power spectrum against two thresholds. For example, in FIG. 18, three peaks exceed the threshold G2, but only two peaks exceed the threshold G3. The angular spread of a single peak determined according to threshold G2 is broader than the angular spread determined according to threshold G3. The ratio of the angular spread of the single peak determined by G3 as compared to the spread determined by G2 is a measure of the sharpness of the peak. Alternatively, the threshold against which the angular power spectrum is measured may be moved adaptively until there are at most two peaks in the angular power spectrum above the threshold to reveal the directions of paths 3 and 5. For example, when two sharp peaks occur in the angular power spectrum and the base station transmits two beams, the base station defines the direction of these peaks (i.e., the two distinct angular directions where the power spectrum exceeds threshold G3) to be the angular directions for paths 3 and 5 (FIG. 1). This is referred to as angle of arrival diversity. The base station points steerable beams, or selects fixed beams to point, along respective paths 3 and 5. Persons skilled in the art will appreciate how to extend angular diversity management to more than two beams.

On some occasions, the angular power spectrum includes three or more angular positions that correspond to respective peaks in the angular power spectrum. When the base station has two beams, the base station selects first and second angular positions from the three or more angular positions either (1) based on the avoidance of angles at which co-channel users are located so as to minimize co-channel interference on a system wide basis, or (2) so as to balance power distribution in amplifiers of the transmit station.

The beam widths in a phased array antenna are generally selectable by controlling an amplitude of elements in the beam steering vector (e.g., vector C of FIG. 7). When the antenna system includes a phased array antenna with controllable beam widths and the spectral peaks are sharp, the base station sets or selects beams to be as narrow as

practical given the antenna system in order to concentrate the transmit power in directions along respective paths 3 and 5. Paths 3 and 5 are expected to have good transmission properties since the spectral power peaks are sharp.

On the other hand, when the angular power spectrum is so diffuse that peaks are weak or not indicated, a general angular window is determined based on the angular extent over which the power spectrum exceeds a threshold (e.g., G2 of FIG. 18) or at least the continuous angular extent needed to cover the peaks where the angular power spectrum exceeds the threshold. In such a case, preferred embodiments of the invention select beams such that the sum of the beam widths for all beams used for down link transmissions approximately equals angular spread AS.

When the antenna system includes a phased array antenna with controllable beam widths but the spectral peaks are not so sharp, the base station first determines the angular spread to be the angular extent of the power spectrum that is greater than a threshold or at least the continuous angular extent needed to cover the peaks where the angular power spectrum exceeds the threshold. Then, the base station sets or selects the beam widths for the beams to approximately cover the angular spread. This is referred to as angular power diversity or beam width management. For example, a two beam base station that seeks to cover the angular spread will select a beam width for both beams to be about half of the angular extent, and the base station points the two beams to substantially cover the angular spread.

Extensions to more beams are straightforward as will be appreciated by persons skilled in the art. For example, when the base station has capability for beamspace time encoding in a four beam base station, a beam width is selected for each beam that is approximately one-fourth of the angular spread. In this way the down link transmission will spatially match the channel. It is advantageous to match the coverage of orthogonal beams to the angular spread of the channel to obtain maximum angular diversity gain. However, usually two to four beams are adequate.

When the base station has an antenna system with plural fixed beams (as with a hex corner reflector antenna) and when the angular power spectrum is diffuse and angular spread AS exceeds the beam width of a single beam, a desirable variant of the invention combines two adjacent beams into a single broader beam (e.g., combine two 60 degree

beams into a single 120 degree beam) to better match the radio channel. In such a case, the two adjacent beams are used as a single broader beam employing the same pilot code or orthogonalizing code. In fixed beam base stations, it is advantageous that the number of beams M that can be generated is large (e.g., M>4, and preferably at least 8) so that high beam resolution can be achieved. When a broader beam is needed to better match the channel, two adjacent beams may be combined.

The present invention fits well in a base station where the antenna system employs digital beam forming techniques in a phase array antenna (e.g., antenna 20 of FIG. 7 and antenna 40 of FIG. 10). With digital beam forming techniques, the apparent number elements in an antenna array (i.e., the apparent aperture dimension) can be electronically adjusted by using zero weighting in some of the elements according to the available angular spread. In this fashion, the beam width can be easily adapted by the base station to match the angular spread. This beam width control operates as an open loop control system.

In an alternative embodiment, beam hopping techniques are employed when the angular power spectrum exceeds the threshold in one large angular extent. A beam hopping technique is a technique that covers the angular spread sequentially. For example, when the transmit beams in any one time slot do not cover the angular spread, the angular spread may be covered during subsequent time slots. Consider an exemplary system that has a two beam base station capable of forming 30 degree beams where the angular spread covers 120 degrees (i.e., the width of four beams). In a beam hopping system, the base station forms two 30 degree beams for transmission during a first time slot so as to cover a first 60 degree sector of the 120 degree angular spread, and forms two other 30 degree beams for transmission during a second time slot so as to cover the remaining 60 degree sector of the 120 degree angular spread.

Beam hopping greatly improves performance in radio environments with large angular spreads. It is known that the down link performance degrades in frequency division duplex cellular radio systems when the angular spread becomes large, due at least in part to the increased angular uncertainty in the optimal selection of directions for transmission. In frequency division duplex systems, the up link directions determined

to have good power transmission capacity (low attenuation) could be in a deep fade for a down link transmission due to the different carrier frequencies.

With a large angular spread in the radio environment, the number of possible directions for down link transmission will be large. Instead of selecting the two best directions, spatial diversity is achieved by sequentially forming down link beams to cover all of the potentially good directions where the angular power spectrum exceeds a threshold. This is particularly important in micro-cells or pico-cells where the angular spread can cover the whole sector or the whole cell.

In a scenario where remote station 2 is fixed or of low mobility, beam hopping has additional advantages over selection of the two strongest directions. When the best two directions are selected as the beam transmit directions for a large number of consecutive bursts, there is considerable penalty (in terms of loss of data) if the selected directions are a wrong choice (e.g., down link in deep fade even though up link is good). However, by hopping the beams over a group of potential directions, the loss of data from any one direction that turns out to be in deep fade will be for only a limited duration (e.g., only one time slot). This angular diversity tends to "whiten" the errors generated by selection of bad transmission directions.

Furthermore, the co-channel interference to other remote stations generated during beam hopping transmissions will tend to be whitened by the spatial spreading of the transmitted signal. Co-channel interference can be particularly troublesome when high data bit rate connections are required since high bit rate connections are achieved with high beam powers. The large amount of beam power involved in the high bit rate connection generates highly colored interference (not uniformly distributed) when a non-hopping scheme is employed by the base station for beam selection.

In FIG.19, another embodiment of the invention includes base station 210 and remote station 230 as described with reference to FIG. 14. In the present embodiment, base station 210 includes weighting amplifiers 102 and 104 to apply respective weights W1 and W2 to respective feed signals CH1 and CH2. In the present embodiment, weights W1 and W2 are complex numbers or at least phase and amplitude pairs to control both the amplitude and phase of the signal transmitted from antennas 16 and 18. The weighted signals may alternatively be transmitted from directional antennas 106 and 108.

FIG. 19 depicts diplexers 16D and 18D coupled between the weighting amplifiers and the respective antennas to duplex the antennas so they may be used in an up link receive mode as well as a down link transmit mode; however, a separate base station antenna may be used to receive up link signals.

In a preferred variant, one antenna is used as a reference with its corresponding weight set to 1+j0 (or amplitude = 1, phase = 0°). The other weight is determined relative to the reference weight. In general, base station 210 may employ two or more channels, each with an antenna, diplexer, weighting amplifier and all associated encoders. If M is the number of transmitting antennas, then the number of weights that must be determined is M - 1 since only differential information (i.e., weights) need to be determined. Without loss of generality, the following description focusses on two transmitting antennas (M = 2) so that only one complex number weight need be determined.

In FIG. 19, remote station 230 includes remote station antenna 232, remote station receiver 234 coupled to remote station antenna 232 through diplexer 233, signal measurement circuit 238, and processor 240. Receiver 234 constitutes a circuit by which remote station 230 receives first and second signals from respective first and second transmit antennas. Signal measurement circuit 238 and processor 240 and control modules described herein constitute a circuit by which remote station 230 determines channel state information based on the received first and second signals and segments the channel state information into a plurality of channel state information segments. Signal measurement circuit 238 measures the signal strength (and phase) received from each of the plural orthogonal antennas, and processor 240 determines channel state information. Signal measurement circuit 238 measures an instantaneous signal strength (and phase) received, or in an alternative variant, measures an averaged signal strength received and a phase at a reference time.

The processor determines the channel state information from information provided by signal measurement circuit 238. The processor selects a reference signal from among the signals received from the different antennas. For each of the plural antennas, the processor divides the received signal strength (and phase) determined by signal measurement circuit 238 by the selected reference signal strength (and phase). This ratio is determined as a ratio of complex numbers (or phase/amplitude pairs). The

ratio for the reference antenna is, by definition, 1 +j0. In the case of two antennas, there is only one ratio to be sent, the ratio of the reference antenna being a constant reference.

Processor 240 determines the channel state information from the normalized ratio or ratios. Each ratio includes both amplitude and angle information. It is the object of this process to adjust the phase of signal transmitted from the two antennas (or more) so that they will constructively reinforce at remote station 230. To ensure constructive reinforcement, it is desired to phase delay or advance a signal transmitted from each antenna relative to the reference antenna. For example, if first antenna 16 is the reference antenna, then the angle portion of the ratio for the signal received from second antenna 18 is further examined. If this angle is advanced 45 degrees relative to the reference antenna, it will be necessary to introduce a 45 degree delay at the transmitter for second antenna 18 to achieve constructive reinforcement at remote station 230. Thus, processor 240 determines the amount of phase delay or advance needed to achieve constructive reinforcement at remote station 230 by adding the desired additional delay to the phase of the initial transmitted signal, and if the addition result is greater than 360, then subtracting 360. This phase angle then becomes the phase angle transmitted as part of the channel state information.

Processor 240 also determines the amplitude part of the channel state information. The object here is to emphasis the antenna with the best path (i.e., lowest attenuation path) from the antenna to remote station 230. The total power transmitted from all antennas may be regarded here as constant. The question to be resolved by the amplitude part of the channel state information is how to divide up the total transmitted power.

To do this, processor 240 measures the channel gain (the inverse of the attenuation) by computing, for each antenna, the ratio of the power received divided by the power received in the reference signal. The power received is the square of the signal strength measured by signal measurement circuit 238 (i.e., $P_i=(a_i)^2$ where a_i is the signal strength from antenna i). The signal transmitted through each different antenna or antenna beam includes its unique and mutually orthogonal pilot code modulated on a signal transmitted at signal power P_{TX}. The remote station measures the complex channel

impulse response, $H_i = a_i \exp(\phi_i)$ as a ratio of the signal received divided by the reference

signal received where ϕ_i is the relative phase of the signal being measured and a_i is the relative signal strength. Then P_i is determined as the square of a_i. The relative channel response for each antenna is measured in terms of received power. If only one bit were reserved in the up link signaling channel for amplitude feedback information, the bit would preferably command 80% of the total power to be transmitted by the antenna with the lowest attenuation path to remote station 230 and command 20% of the total power to be transmitted by the antenna with the highest attenuation path.

If two bits were reserved in the up link signaling channel for amplitude feedback information, the bits could define four amplitude states. For example, processor 240 would compute a ratio between the path attenuation from antenna 16 and the path attenuation from antenna 18 and then slice the ratio according a predetermined range of values that this ratio can take. The slicing process defines four sub-ranges and identifies into which of the four ranges the computed ratio fits. Each sub-range would define the desired split of the total power transmitted by two antennas 16 and antenna 18 to be, for example, 85%/15%, 60%/40%, 40%60% and 15%/85%, respectively. The two bits would thus encode one of these splits as the desired split in the total power transmitted by two antennas.

Persons skilled in the art will appreciate, in light of these teachings, that the amplitude portion of the channel state information may be computed by various means. Described here is a table look up means, but other means to compute the split of the total power to be transmitted are equivalent. It will be appreciated that three or more bits may be used to define the power split.

Processor 240 also segments the channel state information (including the amplitude portion and phase angle portion described above) into a plurality of channel state information segments based on the design . Remote station 230 further includes transmitter 242 to send the plurality of channel state information segments to base station 210.

The channel state information to be transmitted is a complex coefficient in the form of phase and amplitude information, and it is to be transmitted from remote station

230 to base station 210 in a number of segments (N segments) carried in corresponding slots in an up link signaling channel. A partition of the N slots into N1 and N2 (where N = N1 + N2) is done in such a way that the first N1 slots carry phase information and the remaining N2 slots carry amplitude information. In principle N1 and N2 can be arbitrarily chosen, but a common value for these parameters could be N1 = N2 = N/2. Assume that each slot reserves K bits for carrying the corresponding information segment. The the phase can be resolved to an accuracy of:

$$\phi_{\min} = \frac{360}{2^{N_1 K}} ,$$

and the amplitude can be resolved to an accuracy of:

$$A_{\min} = \frac{A_{\max}}{2^{N_2 K}} ,$$

where A_{max} is the maximum amplitude.

For example, assume that the number of slots, N, is 6, and three slots are reserved for each of N1 and N2. Assume that the number of bits per slot, K, is 1, and assume that the maximum amplitude, A_{max}, is 3 volts. Then, the accuracy of the phase and amplitude are $\phi_{min} = 45°$, and the amplitude A_{min} is 0.375 volts. However, if the number of bits per slot, K, were increased to 2, the accuracy of the phase and amplitude that could be sent would be $\phi_{min} = 5.6°$, and the amplitude A_{min} is 0.05 volts.

In general, a quantized or truncated version of the exact channel state information is formed so that the bits in the truncated version exactly matches the number of bits available in the up link signaling channel. The truncated version is segmented into phase segments ϕ_i (i = 1 to N1), and the segments are transmitted in a hierarchal order so that the most significant bit (MSB) is transmitted in the first segment and the least significant bit (LSB) is transmitted in the last segment. Similarly, each amplitude segment, A_i (i = 1 to N2) contains a quantized or truncated segment of the exact channel state information (the ratio) and it is transmitted in a hierarchal order.

The present embodiment of the invention improves the down link performance of mobile communications due to improved phase angle and amplitude accuracy for use

in forming down link beams. This embodiment is particularly suitable for low mobility environments, and it suits high data rate applications in indoor and pedestrian environments. The embodiment is particularly suited for high bit-rate wireless data applications for laptop computers.

For example, assume the remote station is moving at a speed of v = 1 meter per second (3.6 kilometer per hour) and the carrier frequency is 2 gigahertz ($\lambda = 0.15$ meters). The maximum Doppler frequency f_D is v/λ and the channel coherence time T_C is computed to be:

$$T_C = 1/(2f_D) = \lambda/(2v) = 75 \text{ milliseconds.}$$

It can be assumed that the channel state information will remain stable (nearly constant) over a time period equal to $T_C/10$, and therefore, the channel state information may be sent from remote station 230 to base station 210 in during this stable time period of 7.5 milliseconds. Since wideband CDMA (WCDMA) standards define slot durations to be 0.625 milliseconds, one can use 12 slots to send the channel state information back to the base station.

There are several ways to pack the channel state information in the up link slots. Table 1 illustrates an example based on only one bit per slot (K = 1). In Table 1 three-bit accuracy is used for both the phase angle and the amplitude information. The phase angle is transmitted in the first 6 slots, and the amplitude information is transmitted in the last 6 slots. In both cases, the most significant bits are transmitted first. In slot 1, the most significant bit of the three-bit phase angle is transmitted. In slot 2, the same bit is repeated to improve reliability. After that, the remaining phase angle bits are transmitted, and the amplitude information bits are sent in the same fashion. The first bit gives the phase angle to an accuracy of 180° as if in a one-bit . After slot 3, the phase angle is sent to an accuracy of 90° as if in a two-bit , and after slot 5, the phase angle is sent to an accuracy of 45° as in the three-bit . If it is assumed that the phase angle changes about 360° during the coherence time of the channel, then in the above example, the phase angle will change about 36° in the 7.5 millisecond time period it takes to send 12 slots. This corresponds well to the phase accuracy achievable with three-bit data (45°).

After slot 7, the amplitude information is sent to an accuracy of 0.5 of the maximum amplitude as if in a one-bit . After slot 9, the amplitude information is sent to

an accuracy of 0.25 of the maximum amplitude as if in a two-bit , and after slot 11, the amplitude information is sent to an accuracy of 0.125 of the maximum amplitude as in a three-bit .

Table 1

Format For Sending Channel State Information To The Base Station			
Slot Number	Feedback Bit	Slot Number	Feedback Bit
1	Phase MSB	7	Amplitude MSB
2	Phase MSB	8	Amplitude MSB
3	Phase Bit 2	9	Amplitude Bit 2
4	Phase Bit 2	10	Amplitude Bit 2
5	Phase LSB	11	Amplitude LSB
6	Phase LSB	12	Amplitude LSB

In general the phase information is more important than the amplitude information. The optimum maximal ratio combining performs only about 1 dB better than the equal gain combining that would be used if there were no amplitude information feedback, and thus, a larger allocation to phase bits (N1) and a lesser allocation to amplitude bits (N2) has advantages. For example, one could allocate three phase bits and two amplitude bits so that the feed back channel state information could be sent in a WCDMA format without redundancy in 3.125 milliseconds.

The tradeoff between the allowed feedback capacity (e.g., one or more bits/slot), the feedback reliability (e.g., number of repeated or redundant bits) and the feedback accuracy (e.g., number of phase angle and amplitude bits) is application and environment specific. For example, a three-bit check code in a well known SECDED (single error correct, double error detect) format may be appended to 8 bits of information to provide redundancy error checking. Persons of ordinary skill in the art, in light of these teaching, will appreciate how to match the feedback capacity, the feedback reliability and the feedback accuracy to the application and environment.

Processor 240 (FIG. 19) segments the channel state information into a plurality of channel state information segments according to the format defined by the system

modes. In fact, a system may be designed with multiple modes, each mode defining different formats. For example, one mode may send only phase angle correction information commanding equal amplitudes to each of the antennas, and another mode may send three bits of phase angle information and one bit of amplitude information. Then transmitter 242 encodes the plurality of channel state information segments in an up link signaling channel and sends the encoded information through diplexer 233 and antenna 232 to base station 210.

In one variant of the embodiment, there are several modes requiring from 1 to, for example, 20 bits to express the channel state information in the up link signaling channel. In this variant, processor 240 determines the rate at which the channel state information changes based on changes from update to update. When the rate is slow, indicating a slow moving or stationary remote station, the feedback mode is adaptively changed to a mode that permits more data bits of the channel state information to be sent to the base station. However, when the channel state information changes rapidly, indicating that the remote station is rapidly moving, then the feedback mode is adaptively changed to a mode that sends fewer bits for each channel state information update.

Base station 210 receives the information encoded in the up link signaling channel and decodes the plurality of channel state information segments in receiver/detector 220. Processor 220P then reconstructs the channel state information from the received plurality of channel state information segments and produces weights W1 and W2. Weights W1 and W2 are provided to respective amplifiers 102 and 104 to weight first and second feed signals CH1 and CH2 to feed to the respective first and second antennas 16 and 18 based on the reconstructed channel state information.

Two variants of this embodiment may be implemented in processor 220P. First, the processor may collect all segments to reconstruct the total channel state information before forming weights W1 and W2 to apply to amplifiers 102 and 104. Alternatively, the channel state information is sent to the base station phase angle first and within the phase angle segments, most significant bit first. The values of W1 and W2 may be updated within the processor as each bit is received to provide more immediate feedback to amplifiers 102 and 104. This produces, in effect, a higher feedback bandwidth.

In FIG. 20, a method practice on processor 240 includes several steps that are typically implemented in the processor with software modules and/or logic. However, persons skilled in the art will appreciate that the steps may be implemented in the processor using ASIC or other custom circuitry.

In step S2, for each of the plural antennas, the processor receives the received signal strength and phase (a complex number) as determined by signal measurement circuit 238. In step S4, the processor selects one of the received signals to be a reference signal. This selection may be arbitrary or it may be to select the signal with the greatest phase lag (least likely to need to or want to be slowed down). In step S6, the processor divides the received signal strength and phase (a complex number) determined by signal measurement circuit 238 by the received reference signal strength and phase (a complex number). The ratio for the reference antenna is, by definition, 1+j0. In the case of two antennas, there is only one ratio to be determined and sent, the ratio of the reference antenna being a constant reference.

In step S8 (FIG. 20), processor 240 determines the amount of phase delay or advance needed at each transmitting antenna to achieve constructive reinforcement at remote station 230. If the reference signal is chosen to be the signal with the most lag, the remaining signals may achieve phase alignment with the reference signal by adding a delay at the antenna. Step S8 determines the required additional delay, but if the addition phase delay added to the phase of the non-reference signal results in a phase that is greater than 360 degrees, then subtract 360. This phase angle then becomes the phase angle transmitted as part of the channel state information. Persons skilled in the art in light of these teachings will appreciate that step S8 may be performed in the base station so that only the phase angle of the channel impulse response need be sent in the up link signaling channel.

In step S10, power management information to define the transmit distribution (the allocation of the total power among the transmit antennas) is determined. Persons skilled in the art will appreciate in light of these teachings, that the amplitude portion of the channel state information may be computed by various means. Described here is a table look up means, but other means to compute the split of the total power to be transmitted are equivalent.

For example, the relative amplitude and relative phase of the signal from each antenna may be transmitted in the up link signaling channel for the base station to further process. Alternatively, the remote station may determinine in step S10 an indicia of the desired power distribution. If only one bit were reserved in the up link signaling channel for amplitude feedback information, the bit would preferably command 80% of the total power to be transmitted by the antenna with the lowest attenuation path to remote station 230 and command 20% of the total power to be transmitted by the antenna with the highest attenuation path. If two bits were reserved in the up link signaling channel for amplitude feedback information, the bits could define four amplitude sub-ranges. For example, 85%/15%, 60%/40%, 40%60% and 15%/85%, respectively. The two bits would thus encode one of these sub-ranges as the desired split in the total power transmitted by two antennas. Extensions to more antennas or to the use of more bits to represent the amplitude portion of the channel state information will be apparent to persons of ordinary skill in the art. The exact nature of the table look up or other means depends on the number of bits reserved in the up link format to carry the amplitude portion of the channel state information.

In step S12, the channel state information is segmented and packed into the formats described herein (e.g., Table 1). In step S14, the segments are sequentially transmitted in the up link signaling channel to the base station. From there, the respective weights for the antennas are recovered and applied to amplifiers 102 and 104 (FIG. 19).

In Frequency Division Duplexed systems where up link and down link communications are carried out over different frequencies, it is not possible to exactly determine the down link channel state from up link information since the two directions are based on different frequencies. The present system has the advantage of measuring the down link channel state from down link data and then sending commands in the up link signaling channel to adjust the amplitude and phase of the transmitted down link signals.

In FIG. 21, antenna 1 of the base station is a sector coverage type of antenna. Antenna 1 sends a signal to remote station 2 over direct path 3; however, another multi-path signal reflects off of radio wave scatter 4 and travels over multi-path 5. As a result, remote station 2 receives two replicas of the signal at slightly different times. In FIG. 22,

the two replicas are depicted as signals received at time nT and time nT+τ where τ is the additional time delay that occurs due to the additional length of multi-path 5 when compared to direct path 3. The multi-path delay may be such as to cause destructive interferences between the two signals received over the two paths. Additional radio wave scatterers may create even more multi-path signals.

A conventional Rake receiver correlates a local signal (e.g., the spreading code of a CDMA signal) and the received signal that includes signal replicas received with different delays. With correct delays, the signals are coherently combined to reinforce energies. When the local signal (e.g., desired spreading code) is correlated with a signal from a desired signal path, the local signal is also correlated with every one of the other signal replicas (e.g., signal replicas from signal paths with different delays). The terms corresponding to the correlation with the other signal replicas are unwanted terms, and they tend to degrade the performance of the system. The unwanted correlation terms also cause a loss of orthogonality between different users with different codes, and as a result, co-channel users start to interfere with each other. The degradation effect becomes more pronounced with short spreading codes that are typically used in high bit rate links.

The present invention operates the Rake receiver in an unconventional fashion. Using beam forming, the present invention separates different signal paths and applies pre-transmission time shift compensation on each signal replica (e.g., each beam) so that all signal replicas arrive at the receiver simultaneously. In this manner, the receiver appears to receive a signal processed only through a 1-tap channel even though it actually receives and coherently combines multiple signals over multiple paths (e.g., paths 3 and 5 in FIG. 1). This avoids a loss of orthogonality and minimizes or eliminates cross correlation terms that might otherwise degrade system performance.

In an embodiment of the present invention, the desired data is included in two or more space-time coded signals. The signals are identified by unique and mutually orthogonal signature codes. If one of the space-time coded signals is significantly delayed with respect to another, the orthogonality of the signature codes may be reduced. It is preferred to delay the shortest path signal so as to arrive at remote station 2 at the same time that the longer path signal arrives at remote station 2.

In FIG. 23, an exemplary system includes antenna 1 and remote station 2. Exemplary antenna 1 may be a Butler matrix multi-beam antenna array or any other multi-beam antenna array. The desired data in this example are encoded into two space-time coded signals I2 and I5. Space-time coded signals I2 and I5 are transmitted in beams D2 and D5, respectively. Beam D5 sends signal I5 to remote station 2 over direct path 3. Beam D2 sends signal I2 to remote station 2 over indirect multi-path 5.

In FIG. 26, an exemplary encoder for the generation of space-time coded signals I2 and I5 is depicted. FIG. 26 is similar to FIG. 2, except that antennas 16 and 18 of FIG. 2 are replaced by the multi-beam antenna of FIG. 23 and a programmable delay line (e.g., a selectable multi-tap delay line) is coupled between multiplier 14 and the multi-beam antenna. Multiplier 12 encodes signal CH1 with a signature code (OC) that is mutually orthogonal to the signature code that is encoded in the signal CH2 by multiplier 14. The signature codes may be variously orthogonal training sequences, pilot codes or spreading sequences. Using these signature codes, remote station 2 separates the signal that is received in a direct path from beam D5 from the signal that is received in an indirect path from beam D2 as long as the signature codes remain orthogonal. Persons skilled in the art will appreciate that the two beams and corresponding space-time coded signals depicted in FIGS. 23 and 26 may be generalized to more than two and that additional programmable delay lines may be needed to time synchronize all signals.

The direct signal from beam D5 is received at remote station 2 before the indirect signal from beam D2 is received by a time τ as depicted in FIGS 24 and 25. In order to maintain best orthogonality between the signature codes, it is desirable to align the signals in time. A receiver (possibly at the base station and possibly at remote station 2 as discussed below) determines the time delay τ necessary to align the signals. The last signal received at remote station 2 (e.g., signal I2) may be regarded as a reference space-time coded signal. The remaining signals may then be regarded as at least one remaining space-time coded signal (e.g., signal I5). In this embodiment, the at least one remaining space-time coded signal is delayed in the programmable delay line of the base station (see FIG. 26) before being transmitted. The signal or signals is or are delayed by a sufficient delay to ensure that each of the at least one remaining space-time coded signal will align in time with the reference signal when received at the remote station. In the example

depicted in FIG. 23, the last signal received at remote station 2 is signal I2 due to the extended length of multi-path 5. Signal I5 will need to be delayed so that it will arrived at remote station 2 at the same time that signal I2 arrives at remote station 2.

In both the space-time diversity technology (FIG. 2) and the beam-space diversity (FIG. 23), it is important for the remote receiver to separate signals CH1 and CH2 as discussed above. This is achieved by using orthogonal signature codes in various forms. The difference in time of arrival when the signals from the two paths, direct path 3 and multi-path 5, arrive at remote station 2 is referred to as the delay spread. When the delay spread does not exist or is minimal, the orthogonality of the signature codes is preserved. However, in frequency selective channels where there exists a considerable delay spread of the signature codes, the orthogonality between the channels may be lost, and remote station 2 will find it difficult to separate signals carried in the respective channels. Most common coding sequences are characterized by non-ideal cross-correlation functions (CCFs) which have a low or zero value only for a given phase relationship between the signature codes, and for other phase relationships, the CCFs are non-zero.

Plural space-time diversity signals intended for transmission to remote station 2 over multi-path channels will undergo different delays. Because the value of the CCF at a given out-of-phase position is typically non-zero and different from position to position, the effect of different path delays imposed by the radio channels on the transmitted signals will be to diminish the orthogonality between the signature codes used by remote station 2 to separate the signals. This loss of orthogonality results in a deterioration in the diversity gain that would otherwise be achieved by the space-time code transmission of signals between a base station and a remote station in a wireless communication system.

In the present embodiment, a multi-beam antenna array associated with the base station receives an up link signal from the remote station of interest in each of the plural beams of the multi-beam antenna array. The up link signal may be a pilot signal, an up link signaling channel, or any other up link channel that identifies the source of the signal as the remote station of interest. The up link signal is received as plural signals derived from radio signals received in corresponding plural beams of the multi-beam antenna array.

For each of the plural received signals, a receiver at the base station separates a signal component identified by a signature code as originating at the particular remote station of interest. The received signal component of each of the plural beams includes a replica of the identified signal for the particular remote station of interest at a particular time delay or delay spread relative to the signal component of a reference beam. A receiver at the base station processes the plural signal components from their respective beams to identify a reference beam as containing the last received signal component and a delay spread needed to align each of the other signal components received from their respective beams with the signal component received in the reference beam. When the base station serves more than one remote station, this process can be repeated for each remote station or for selected remote stations. The selected remote stations could be those with high transmit power. High transmit power might be required by, for example, high data rate requirements.

FIG. 27 depicts a representative channel impulse response or delay distribution profile 300 for a 16 beam base station system that is similar to the 8 beam base station system depicted in FIG. 23. The base station measures the delay spreads τ associated with each beam of the multi-beam antenna. For signals received that have signal strengths above a threshold, an "x" indicates instantaneous and/or averaged signal strength exceeding a given threshold. Directions D3, D6 and D12 depicted at 304, 306 and 308 respectively, include signals with a minimum delay spread (e.g., spanning delays τ_4 through τ_6). If several potential directions are available, preferred directions among the available directions are selected based on additional criteria, such as the whitening of generated interference, the even distribution of power in the plurality of power amplifiers used by the base station and the avoidance of directions where greater than average interference could be caused to co-channel users. For example, a high power beam could cause interference to one or many low bit rate users if the low bit rate users are located within the area illuminated by the high power beam. In some favorable situations, beam hopping can also be applied in order to achieve more effective interference whitening.

In operation, the base station selects directions having minimal delay spreads. For example, the base station selects at least two beams of plural beams that may be formed

by the multi-beam antenna array for transmission of at least two space-time coded signals in corresponding beams of the at least two beams. The at least two beams include a reference beam and at least one remaining beam. The base station also determines from delay distribution profile 300 a time delay corresponding to each beam of the at least one remaining beam for use in programming the programmable delay line.

The base station encodes each signal of the at least two space-time coded signals with a signature code that is mutually orthogonal to each other signature code encoded in the at least two space-time coded signals so as to form a reference space-time coded signal and at least one remaining space-time coded signal (see 12 and 14 of FIG. 26). In the example of FIG. 23, the reference space-time coded signal may be regarded as signal I2 and the at least one remaining space-time coded signal may be regarded as signal I5. However, persons skilled in the art will appreciate in light of these teaching how to extend the present embodiment to more than two space-time coded signals.

The base station delays each signal of the at least one remaining space-time coded signal to form at least one delayed space-time coded signal (e.g., signal I5 in FIG. 26). The base station then transmits the reference space-time coded signal (e.g., signal I2) and the at least one delayed space-time coded signal (e.g., signal I5) in respective beams of the at least two beams so that both the reference space-time coded signal and the at least one remaining space-time coded signal arrive at remote station 2 at the same time.

The present embodiment does not rely on a feedback channel from the remote station to the base station. Instead, directions of transmission are selected by the base station solely from up link measurements of normal signaling signals. By averaging the up link channel response over a long time to mitigate fast fading, the power response of the down link channel response can be estimated. The indicated up link and down link channels are reciprocal in the power sense.

However, in frequency division duplex (FDD) systems, a feedback measurement could provided improved results at the cost of additional complexity. In frequency division duplexed systems where up link and down link communications are carried out over different frequencies, it is not possible to exactly determine the down link channel state from up link information since the two directions are based on different frequencies.

The just described embodiment describes an embodiment where the base station measures the up link channel response as a surrogate for the down link channel response. To obtain the complete down link channel impulse response, it is necessary to measure the down link channel directly, and send the down link channel information in a feedback channel from the remote station that does the measuring to the base station that needs the measurements (e.g., delay distribution profile 300).

Rather than performing the calculation required for direction selection and delay in the base station, the remote station participates in or performs these functions. An agreed upon standard signal is sent from the base stations to all remote stations with an identifier or signature coded encoded in each beam, such as mutually orthogonal pilot or training sequences or spreading codes. The remote station would then measure the channel impulse response (e.g., delay distribution profile 300) and inform the base station of the preferred directions and delays for transmission.

Persons skilled in the art will appreciate in light of these teachings that the channel performance may be measured in a two step process. In the first step, the base station makes an estimate of the up link channel's impulse response and uses this estimate as a surrogate for the down link channel's impulse response. Then, the base station applies the delays to the at least one remaining space-time coded signal that are indicated by the first estimate process.

In the second step, the down link channel is measured directly. An agreed upon standard signal is sent from the base station to all remote stations with an identifier or signature coded encoded in each beam, such as mutually orthogonal pilot or training sequences or spreading codes. The remote station would then measure the channel impulse response (e.g., delay distribution profile 300) and inform the base station over a feedback channel of the preferred directions and delays for transmission.

In FIG. 28, set up process S200 measures the up link channel response and sets the measured delays to control the down link channel transmission. Process S200 includes step S202 to measure the channel response, step S204 to select beams to use, step S206 to determine time delays for the selected beams, and step S208 to configure variable delay lines in the base station (see FIG. 26) to impose the determined delays. The variable delay lines may be constructed from a sequence of fixed delay elements with

multiple taps disposed between the elements. The delay line is varied by selecting different taps as an output using a switch. In step S204, the base station selects at least two beams of plural beams formed by a multi-beam antenna array associated with a base station (although only two beams are shown in FIGS. 23 and 26). In the beams are transmitted corresponding at least two space-time coded signals produced by a space-time encoder (although only two signals are shown in FIGS. 23 and 26). The at least two beams include a reference beam and at least one remaining beam. In step S206, the base station determines a time delay corresponding to each beam of the at least one remaining beam. In step S208, the base station sets into a variable delay line the time delay corresponding to each beam of the at least one remaining beam. Each variable delay line is coupled between the multi-beam antenna array and the space-time encoder (see FIG. 26).

In FIG. 29, time align process S220 marks the space-time coded signal for each selected beam with a signature code orthogonal to all other beams in step S222, delays selected beams according to determined delay spreads in step S224 and transmits the delayed signals to the base station in step S226. In step S222, the base station encodes each signal of the at least two space-time coded signals with a signature code that is mutually orthogonal to each other signature code encoded in the at least two space-time coded signals so as to form a reference space-time coded signal and at least one remaining space-time coded signal. In step S224, the base station delays each signal of the at least one remaining space-time coded signal in a respective variable delay line to form at least one delayed space-time coded signal. In step S226, the base station transmits the reference space-time coded signal and the at least one delayed space-time coded signal in respective beams of the at least two beams.

In FIG. 30, a remote station using feedback process S240 measures down link complex channel state information and feeds this information back to the base station. Process S240 includes step S242 to receive at least two identifier signatures (e.g., different pilot signals) from an antenna system associated with a base station, step S244 to determine complex channel state information based on the received signals, step S246 to segment the complex channel state information into a plurality of channel state information segments, and step S248 to send the plurality of channel state information

segments in a sequence to the base station. The sequence of segments sends the most significant bits of the phase angle before the least significant bits of the phase angle. The sequence of segments sends the most significant bits of the amplitude before the least significant bits of the amplitude. The sequence of segments sends a bit of the phase angle before a corresponding bit of amplitude having the same level of bit significance. It is noted that for feedback of the channel impulse response measurements, each beam (or antenna) should be associated with a unique pilot signature that is orthogonal to all other pilot signatures.

It will be appreciated by persons skilled in the art in light of these teachings that various system components may be implemented in electrical circuitry, special application specific integrated circuits (ASICs) or computers or processors that executed software programs or use data tables. For example, encoder 10, multipliers 12, 14 and amplifiers 102, 104 of FIG. 4, 5, 11 or 12 may be implemented in circuitry or ASICs or in some cases, software controlled processors, depending on performance requirements. Beam former 40 of FIG. 11 is typically implemented in circuitry or ASICs and modulators 101,103 and multiplexers 105, 107 are typically implemented in circuitry or ASICs but may be implemented in software controlled processors. Various base station components 212, 214, 216, 218, 220 and 222 and various remote station components 232, 234, 238, 240 and 242 of FIG. 14 may be implemented in circuitry or ASICs but may be implemented in software controlled processors. Various base station components 16D, 18D, 102, 104, 220 and 220P and various remote station components 232, 233, 234, 238, 240 and 242 of FIG. 19 may be implemented in circuitry or ASICs but may be implemented in software controlled processors. It will be appreciated by persons skilled in the art that the various functions described herein may be implemented in circuitry, ASICs or in software controlled processors as the performance requirement dictate.

Having described preferred embodiments of a novel closed loop feedback system for improved down link performance (which are intended to be illustrative and not limiting), it is noted that modifications and variations can be made by persons skilled in the art in light of the above teachings. It is therefore to be understood that changes may be made in the particular embodiments of the invention disclosed which are within the scope and spirit of the invention as defined by the appended claims.

Having thus described the invention with the details and particularity required by the patent laws, what is claimed and desired protected by Letters Patent is set forth in the appended claims.

What is claimed is:

1. A method comprising steps of:

receiving at least two space-time coded signals from an antenna system associated with a first station;

determining complex channel state information based on the received space-time coded signals; and

sending the complex channel state information to the first station.

2. The method of claim 1, further comprising a step of segmenting the complex channel state information into a plurality of channel state information segments, wherein the step of sending the complex channel state information includes sending the plurality of channel state information segments in a sequence.

3. The method of claim 2, wherein the step of segmenting the channel state information includes:

determining a number of phase bits allocated for phase information according to a mode of operation;

rounding and truncating a correction phase angle to fit into the number of phase bits;

determining a number of amplitude bits allocated for amplitude information according to the mode of operation; and

rounding and truncating a correction amplitude according to the number of amplitude bits.

4. The method of claim 2, wherein the step of sending the plurality of channel state information segments includes sending a correction phase angle most significant bit before sending a correction amplitude most significant bit.

5. The method of claim 2, wherein the step of sending the plurality of channel state information segments includes sending a correction phase angle most significant bit before sending a correction phase angle least significant bit.

6. The method of claim 2, further comprising steps of:

receiving the plurality of channel state information segments;

reconstructing the complex channel state information from the received plurality of channel state information segments; and

weighting first and second feed signals to feed respective first and second antennas based on the reconstructed complex channel state information.

7. The method of claim 2, wherein the step of sending includes sequentially sending the plurality of channel state information segments over a time period based on a channel coherence time.

8. The method of claim 1, wherein:

the antenna system includes a multi-beam antenna array;

the step of receiving receives first and second space-time coded signals from respective first and second beams of the multi-beam antenna array; and

the step of determining determines the complex channel state information based on the received first and second space-time coded signals.

9. The method of claim 8, further comprising steps of:

determining by the first station an angular power spectrum of a signal from a second station, the angular power spectrum defining first and second peaks at respective first and second angular positions; and

transmitting the first and second space-time coded signals in the respective first and second beams so that the first and second beams are pointed toward the respective first and second angular positions.

10. The method of claim 1, wherein the antenna system includes a multi-beam antenna array, the method further including steps of:

transmitting the at least two space-time coded signals in respective beams of the multi-beam antenna array with a signature code encoded in each respective signal of the at least two space-time coded signals, the signature codes being substantially

orthogonal so that a second station can separate and measure a channel impulse response corresponding to each space-time coded signal;

measuring the channel impulse response for each space-time coded signal at the second station, the space-time coded signals including a selected set of least attenuated signals and a remaining set of most attenuated signals; and

sending an indicia of the selected set of least attenuated signals from the second station to the first station.

11. The method of claim 1, wherein:

the antenna system includes first and second diversity antennas, the first and second diversity antennas being one of first and second orthogonally polarized antennas and first and second antennas spatially separated by at least one wavelength;

the step of receiving receives first and second space-time coded signals from respective first and second diversity antennas; and

the step of determining determines the complex channel state information based on the received first and second space-time coded signals.

12. The method of claim 1, wherein the antenna system includes plural diversity antennas spatially separated from each other by at least one wavelength, the method further including steps of:

transmitting the at least two space-time coded signals in respective antennas of the plural diversity antennas with a signature code embedded in each respective space-time coded signal, the signature codes being substantially orthogonal so that a second station can separate and measure a channel impulse response corresponding to each space-time coded signal;

measuring the channel impulse response for each space-time coded signal at the second station, the space-time coded signals including a selected set of least attenuated signals and a remaining set of most attenuated signals; and

sending an indicia of the selected set of least attenuated signals from the second station to the first station.

13. The method of claim 1, wherein the antenna system includes first and second diversity antennas, the first diversity antenna being orthogonally polarized with respect to the second diversity antenna, the method further including steps of:

transmitting first and second space-time coded signals in respective first and second diversity antennas with first and second signature codes embedded in the respective first and second space-time coded signals, the first and second signature codes being substantially orthogonal so that a second station can separate and measure a channel impulse response corresponding to each of the first and second space-time coded signals;

measuring the channel impulse response for each of the first and second space time coded signals at the second station, the first and second space-time coded signals including a least attenuated signal and a most attenuated signal; and

sending an indicia of the least attenuated signal from the second station to the first station.

14. The method of claim 1, further comprising a step of transmitting the first and second space-time coded signals with first and second signature codes embedded in the respective first and second space-time coded signals, the first and second signature codes being substantially orthogonal so that a second station can separate a composite signal into the first and second space-time coded signals, wherein the step of receiving receives the first and second space-time coded signals as the composite signal at the second station.

15. The method of claim 1, wherein the complex channel state information includes at least one weight, each weight including amplitude and phase angle information.

16. The method of claim 1, wherein the step of determining complex channel state information includes determining a correction phase angle to adjust a first phase of a first space-time coded signal transmitted from a first antenna relative to a second phase

of a second space-time coded signal transmitted from a second antenna so that the first and second space-time coded signals constructively reinforce at a second station.

17. The method of claim 16, wherein the step of determining a correction phase angle includes:

measuring a first phase angle defined by the first phase;

measuring a second phase angle defined by the second phase; and

determining the correction phase angle defined to be a difference between the second phase angle and the first phase angle.

18. A method comprising steps of:

transmitting at least two space-time coded signals in respective beams of a multi-beam antenna array associated with a first station, the beams transmitting a signature code embedded in each respective space-time coded signal, the signature codes being orthogonal so that a second station can separate and measure a channel impulse response corresponding to each space-time coded signal;

measuring the channel impulse response for each space-time coded signal at the second station, the space-time coded signals including a selected set of least attenuated signals and a remaining set of most attenuated signals; and

sending an indicia of the selected set of least attenuated signals from the second station to the first station.

19. The method of claim 18, further including steps of:

selecting at least two beams for transmission from the first station based on the indicia received from the second station;

transmitting the at least two space-time coded signals in the selected at least two beams;

determining complex channel state information based on the received space-time coded signals; and

sending the complex channel state information to the first station.

20. A method comprising steps of:

selecting at least two beams of plural beams formed by a multi-beam antenna array associated with a first station for transmission of a corresponding at least two space-time coded signals produced by a space-time encoder;

determining a time delay associated with each of the at least two space-time coded signals as received in each respective beam; and

setting into a variable delay line the time delay corresponding to each beam, each variable delay line being coupled between the multi-beam antenna array and the space-time encoder.

21. The method of claim 20, further comprising a step of measuring a channel response based on an up link signal from a second station.

22. The method of claim 21, wherein:

the step of measuring includes receiving plural signal components of the up link signal in corresponding plural beams of the multi-beam antenna array;

the step of selecting includes selecting the at least two beams based on the received plural signal components; and

the step of determining includes determining delay spreads for each of the received plural signal components and assigning to each beam the determined delay spread as the time delay to be set into the respective variable delay line.

23. The method of claim 20, further comprising steps of:

encoding each signal of the at least two space-time coded signals with a signature code that is mutually orthogonal to each other signature code encoded in the at least two space-time coded signals so as to form a reference space-time coded signal and at least one remaining space-time coded signal, wherein the step of setting includes delaying each signal of the at least one remaining space-time coded signal in a respective variable delay line to form at least one delayed space-time coded signal; and

transmitting the reference space-time coded signal and the at least one delayed space-time coded signal in respective beams of the at least two beams.

24. The method of claim 23, further comprising steps of:

receiving the reference space-time coded signal and the at least one delayed space-time coded signal from the multi-beam antenna array;

determining complex channel state information based on the received reference space-time coded signal and the received at least one delayed space-time coded signal; and

sending the complex channel state information to the first station.

25. The method of claim 24, further comprising a step of segmenting the complex channel state information into a plurality of channel state information segments, wherein the step of sending the complex channel state information includes sending the plurality of channel state information segments in a sequence.

26. A system comprising a remote station, the remote station including:

a receiver to receive at least two space-time coded signals from an antenna system;

a processor to determine complex channel state information from the received space-time coded signals; and

a transmitter to send the complex channel state information to a base station.

27. The system of claim 26, wherein:

the processor includes a processor module to segment the complex channel state information into a plurality of channel state information segments; and

the transmitter includes circuitry to send the complex channel state information in a sequence of the channel state information segments.

28. The system of claim 27, wherein the processor module to segment the channel state information includes:

logic to determine a number of phase bits allocated for phase information according to a mode of operation;

logic to round and truncate a correction phase angle to fit into the number of phase bits;

logic to determine a number of amplitude bits allocated for amplitude information according to the mode of operation; and

logic to round and truncate a correction amplitude according to the number of amplitude bits.

29. The system of claim 27, wherein the circuitry to send of the transmitter sends a correction phase angle most significant bit before sending a correction amplitude most significant bit.

30. The system of claim 27, wherein the circuitry to send of the transmitter sends a correction phase angle most significant bit before sending a correction phase angle least significant bit.

31. The system of claim 27, further comprising the base station wherein:

the base station includes a receiver to receive the plurality of channel state information segments;

the base station further includes a processor to reconstruct the complex channel state information from the received plurality of channel state information segments; and

the processor of the base station includes circuitry to weight first and second feed signals to feed respective first and second antennas based on the reconstructed complex channel state information.

32. The system of claim 27, wherein the circuitry to send of the transmitter sequentially sends the plurality of channel state information segments over a time period based on a channel coherence time.

33. The system of claim 26, wherein:

the antenna system includes a multi-beam antenna array;

the receiver receives first and second space-time coded signals from respective first and second beams of the multi-beam antenna array; and

the processor determines the complex channel state information based on the received first and second space-time coded signals.

34. The system of claim 33, further comprising the base station wherein the base station includes:

the multi-beam antenna array;

circuitry to determine an angular power spectrum of a signal transmitted from the remote station, the angular power spectrum defining first and second peaks at respective first and second angular positions; and

circuitry to transmit the first and second space-time coded signals in the respective first and second beams of the multi-beam antenna array so that the first and second beams are pointed toward the respective first and second angular positions.

35. The system of claim 26, further comprising the base station wherein:

the base station includes the antenna system, the antenna system being a multi-beam antenna array;

the base station includes circuitry to transmit the at least two space-time coded signals in respective beams of the multi-beam antenna array with a signature code encoded in each respective signal of the at least two space-time coded signals, the signature codes being substantially orthogonal so that a remote station can separate and measure a channel impulse response corresponding to each space-time coded signal;

the remote station includes circuitry to measure the channel impulse response for each space-time coded signal at the remote station, the space-time coded signals including a selected set of least attenuated signals and a remaining set of most attenuated signals; and

the remote station transmitter sends an indicia of the selected set of least attenuated signals from the remote station to the base station.

36. The system of claim 26, further comprising the base station wherein:

the base station includes the antenna system, the antenna system including first and second diversity antennas, the first and second diversity antennas being one of first and second orthogonally polarized antennas and first and second antennas spatially separated by at least one wavelength;

the receiver receives first and second space-time coded signals from respective first and second diversity antennas; and

the processor determines the complex channel state information based on the received first and second space-time coded signals.

37. The system of claim 26, further comprising the base station wherein:

the base station includes the antenna system, the antenna system including plural diversity antennas spatially separated from each other by at least one wavelength;

the base station further includes circuitry to transmit the at least two space-time coded signals in respective antennas of the plural diversity antennas with a signature code embedded in each respective space-time coded signal, the signature codes being substantially orthogonal so that the remote station can separate and measure a channel impulse response corresponding to each space-time coded signal;

the remote station includes circuitry to measure the channel impulse response for each space-time coded signal at the remote station, the space-time coded signals including a selected set of least attenuated signals and a remaining set of most attenuated signals; and

the transmitter of the remote station includes circuitry to send an indicia of the selected set of least attenuated signals from the remote station to the base station.

38. The system of claim 26, further comprising the base station wherein:

the base station includes the antenna system, the antenna system including first and second diversity antennas, the first diversity antenna being orthogonally polarized with respect to the second diversity antenna;

the base station further includes circuitry to transmit first and second space-time coded signals in respective first and second diversity antennas with first and second signature codes embedded in the respective first and second space-time coded

signals, the first and second signature codes being substantially orthogonal so that the remote station can separate and measure a channel impulse response corresponding to each of the first and second space-time coded signals;

the remote station includes circuitry to measure the channel impulse response for each of the first and second space-time coded signals at the remote station, the first and second space-time coded signals including a least attenuated signal and a most attenuated signal; and

the transmitter of the remote station includes circuitry to send an indicia of the least attenuated signal from the remote station to the base station.

39. The system of claim 26, further comprising the base station wherein:

the base station includes the antenna system and a transmitter coupled to the antenna system, the transmitter of the base station transmitting the first and second space-time coded signals through the antenna system with first and second signature codes embedded in the respective first and second space-time coded signals, the first and second signature codes being substantially orthogonal so that the remote station can separate a composite signal into the first and second space-time coded signals; and

the receiver of the remote station includes circuitry to receive the first and second space-time coded signals as the composite signal.

40. The system of claim 26, wherein the complex channel state information includes at least one weight, each weight including phase angle information.

41. The system of claim 26, wherein:

the antenna system includes first and second antennas; and

the processor to determine complex channel state information includes circuitry to determine a correction phase angle to adjust a first phase of a first space-time coded signal transmitted from the first antenna relative to a second phase of a second space-time coded signal transmitted from the second antenna so that the first and second space-time coded signals constructively reinforce at the remote station.

42. The system of claim 41, wherein the circuitry to determine a correction phase angle includes:

logic to measure a first phase angle defined by the first phase;

logic to measure a second phase angle defined by the second phase; and

logic to determine the correction phase angle defined to be a difference between the second phase angle and the first phase angle.

43. A system comprising a base station and a remote station wherein:

the base station includes a multi-beam antenna array and a transmitter to transmit at least two space-time coded signals in respective beams of the multi-beam antenna array, the beams transmitting a signature code embedded in each respective space-time coded signal, the signature codes being substantially orthogonal so that the remote station can separate and measure a channel impulse response corresponding to each space-time coded signal;

the remote station includes a receiver and a processor to measure the channel impulse response for each space-time coded signal, the space-time coded signals including a selected set of least attenuated signals and a remaining set of most attenuated signals; and

the remote station further includes a transmitter to send an indicia of the selected set of least attenuated signals from the remote station to the base station.

44. The system of claim 43, wherein:

the base station includes a processor to select at least two beams for transmission from the base station based on the indicia received from the remote station;

the base station further includes circuitry to transmit the at least two space-time coded signals in the selected at least two beams;

the processor of the remote station includes circuitry to determine complex channel state information based on the received space-time coded signals; and

the transmitter of the remote station includes circuitry to send the complex channel state information to the base station.

45. A system comprising a base station, the base station including:

a multi-beam antenna array;

first circuitry to select at least two beams of plural beams formed by the multi-beam antenna array for transmission of a corresponding at least two space-time coded signals produced by a space-time encoder;

second circuitry to determine a time delay associated with each of the at least two space-time coded signals as received in each respective beam;

at least two variable delay lines, each variable delay line being coupled between the multi-beam antenna array and a space-time encoder; and

third circuitry to set the time delay corresponding to each beam of the at least two beams into a corresponding delay line of the at least two variable delay lines.

46. The system of claim 45, wherein the first circuitry includes logic to measure a channel response based on an up link signal from a remote station.

47. The system of claim 46, wherein:

the up link signal includes plural signal components, each signal component being a received signal component in a corresponding beam of the plural beams of the multi-beam antenna array;

the logic to measure the channel response includes logic to receive the plural signal components of the up link signal and logic to select the at least two beams based on the received plural signal components; and

the second circuitry includes logic to determine delay spreads for each of the received plural signal components and logic to assign to each beam the determined delay spread as the time delay to be set into the respective variable delay line.

48. The system of claim 45, wherein:

the base station further includes the space-time encoder;

the space-time encoder encodes each signal of the at least two space-time coded signals with a signature code that is mutually orthogonal to each other signature

code encoded in the at least two space-time coded signals so as to form a reference space-time coded signal and at least one remaining space-time coded signal; and

the at least one variable delay line delays each respective signal of the at least one remaining space-time coded signal in a respective variable delay line to form at least one delayed space-time coded signal, the base station transmitting the reference space-time coded signal and the at least one delayed space-time coded signal in respective beams of the at least two beams.

49. The system of claim 48, further comprising a remote station, the remote station including:

a receiver to receive the reference space-time coded signal and the at least one delayed space-time coded signal from the multi-beam antenna array;

a processor to determine complex channel state information based on the received reference space-time coded signal and the received at least one delayed space-time coded signal; and

a transmitter to send the complex channel state information to the base station.

50. The system of claim 49, wherein:

the processor includes a processor module to segment the complex channel state information into a plurality of channel state information segments; and

the transmitter includes circuitry to send the complex channel state information in a sequence of the channel state information segments.

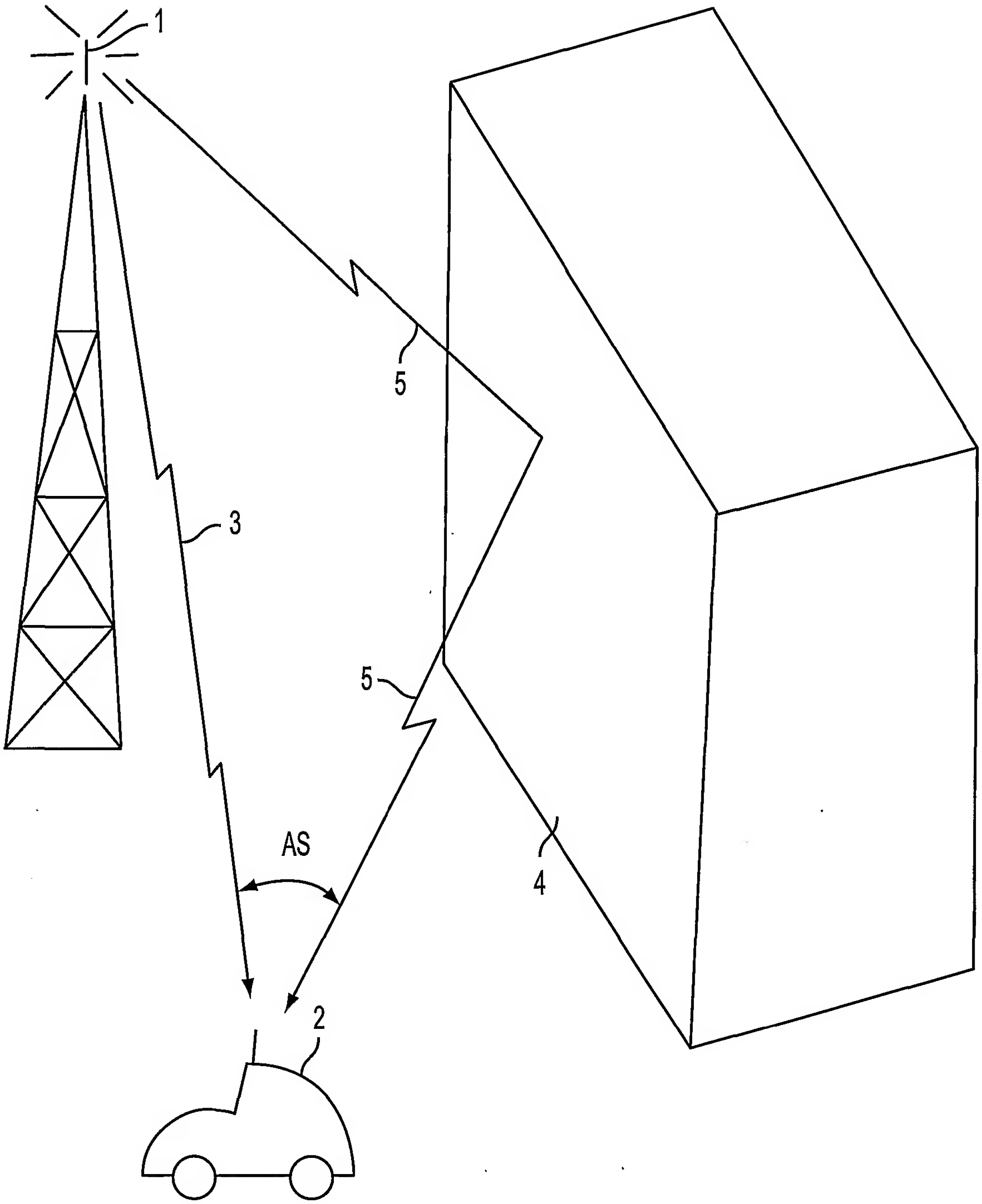


FIG. 1

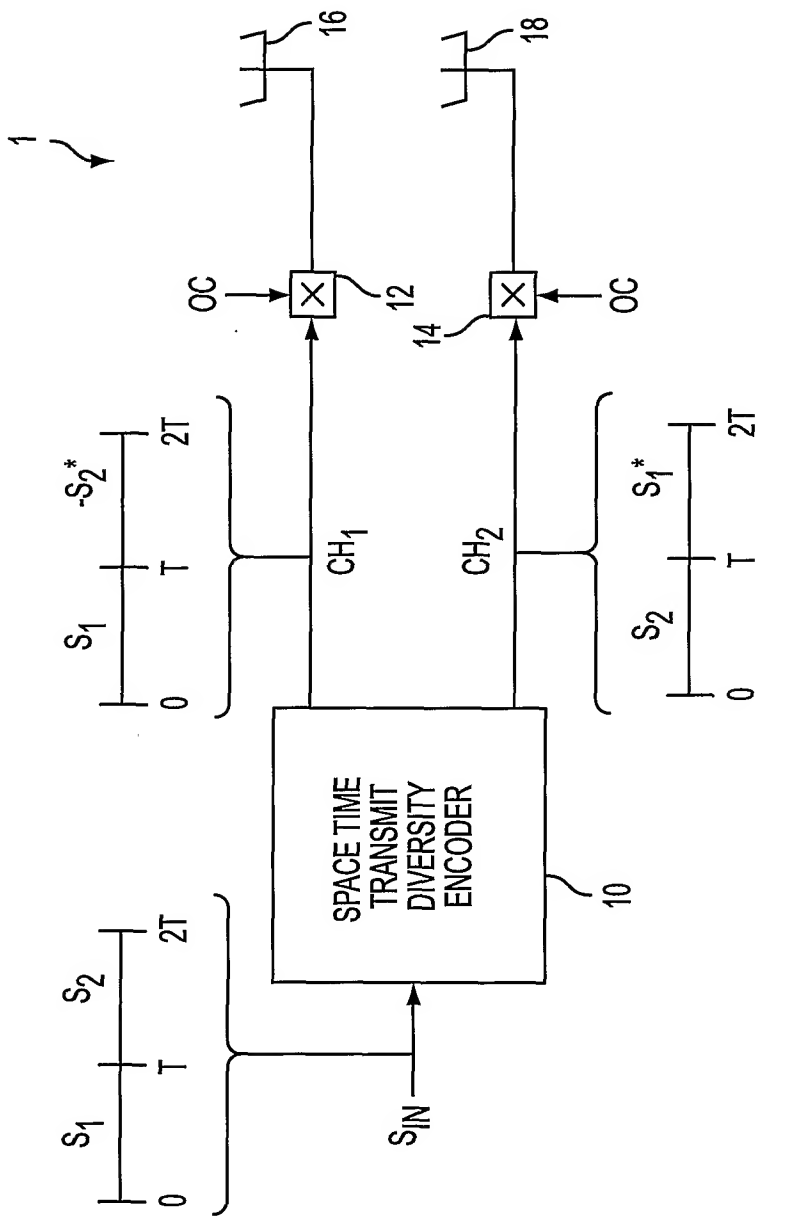


FIG. 2
(PRIOR ART)

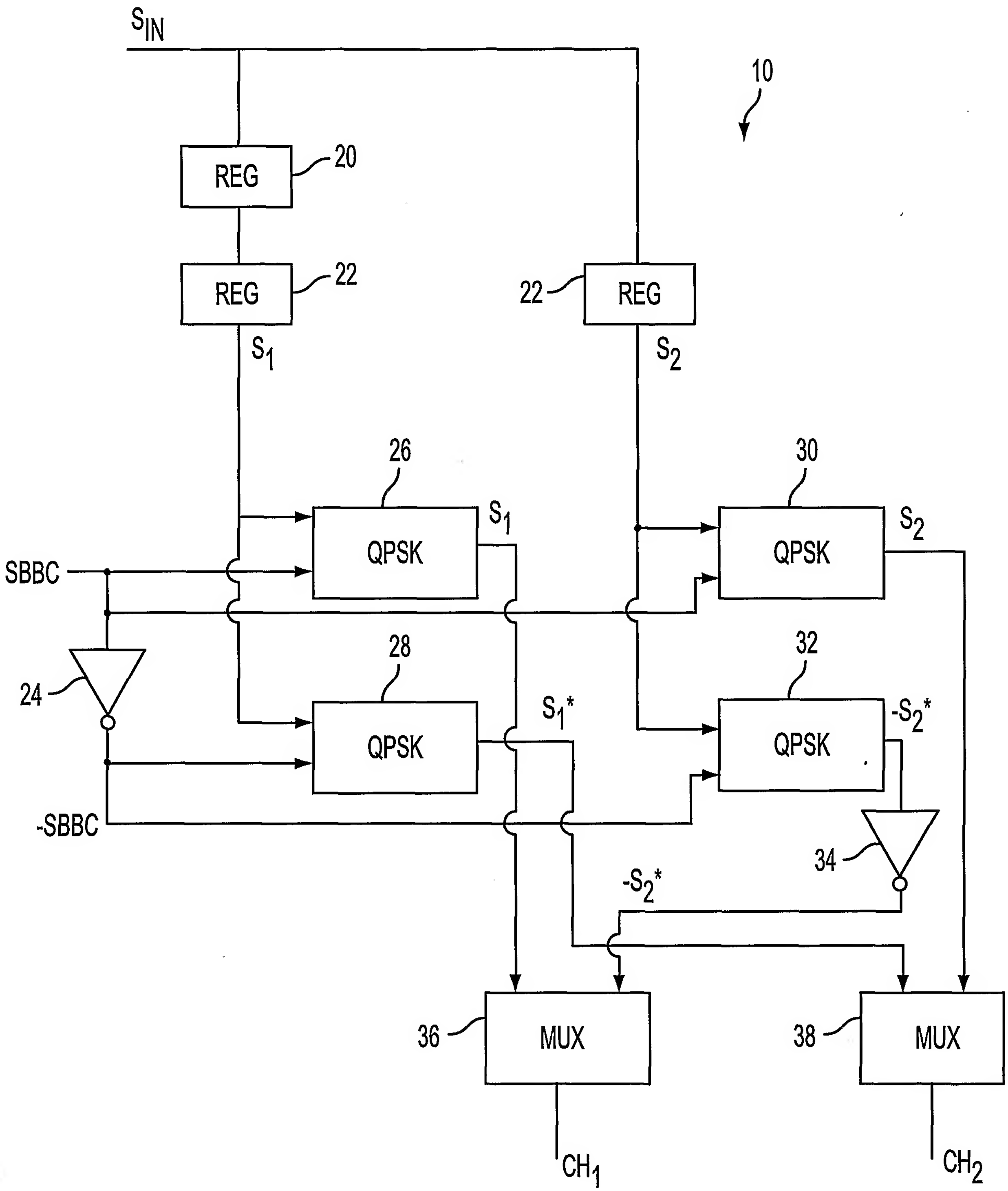


FIG. 3
(PRIOR ART)

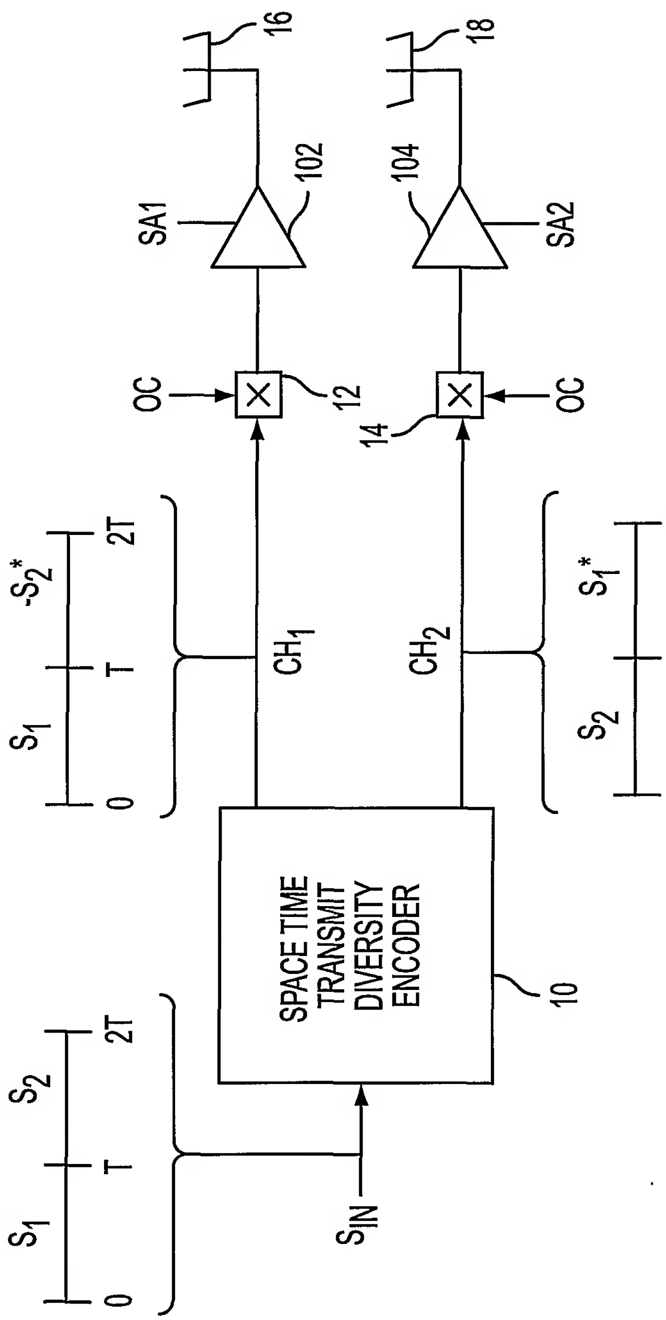

SPACE TIME TRANSMIT DIVERSITY ENCODER
10
S_{IN}
S_1
S_2
0
T
2T
CH_1
CH_2
$-S_2^*$
S_1^*
OC
12
14
SA1
SA2
102
104
16
18

FIG. 4

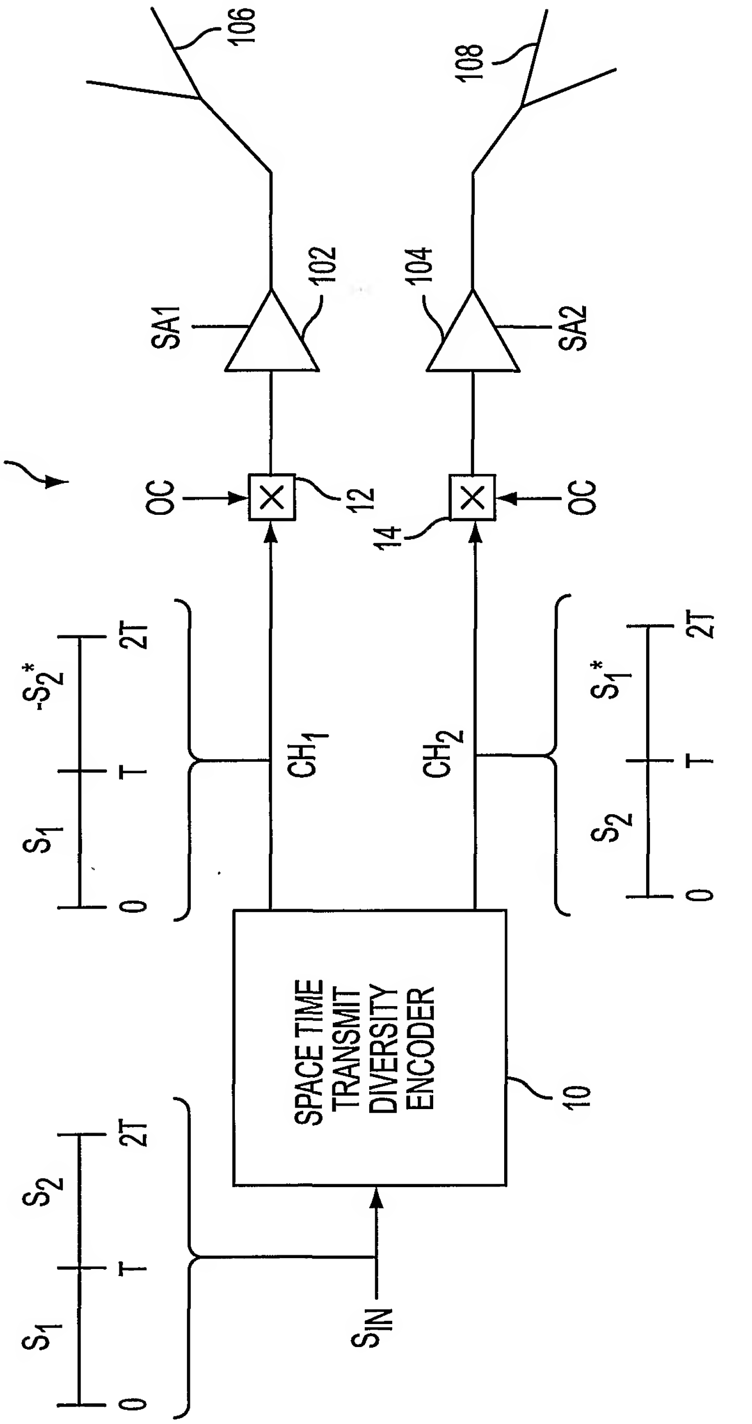

100
SPACE TIME TRANSMIT DIVERSITY ENCODER
10
S_{IN}
S_1
S_2
0
T
2T
S_1
$-S_2^*$
CH_1
CH_2
S_2
S_1^*
OC
12
14
SA1
SA2
102
104
106
108

FIG. 5

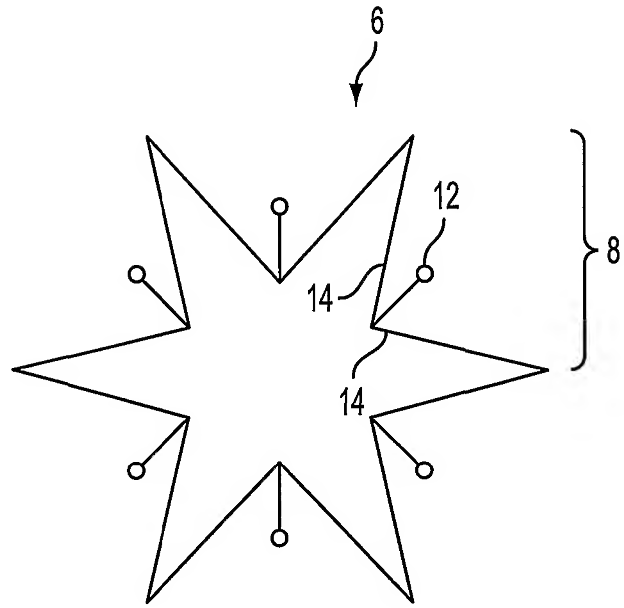


FIG. 6
(PRIOR ART)

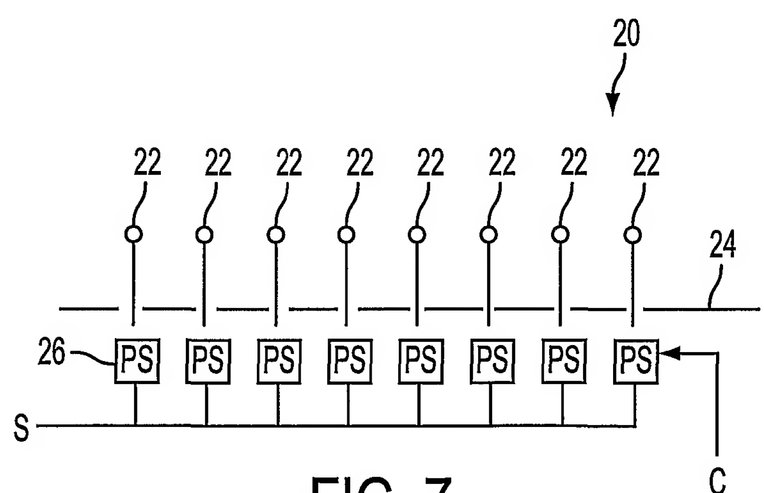


FIG. 7
(PRIOR ART)

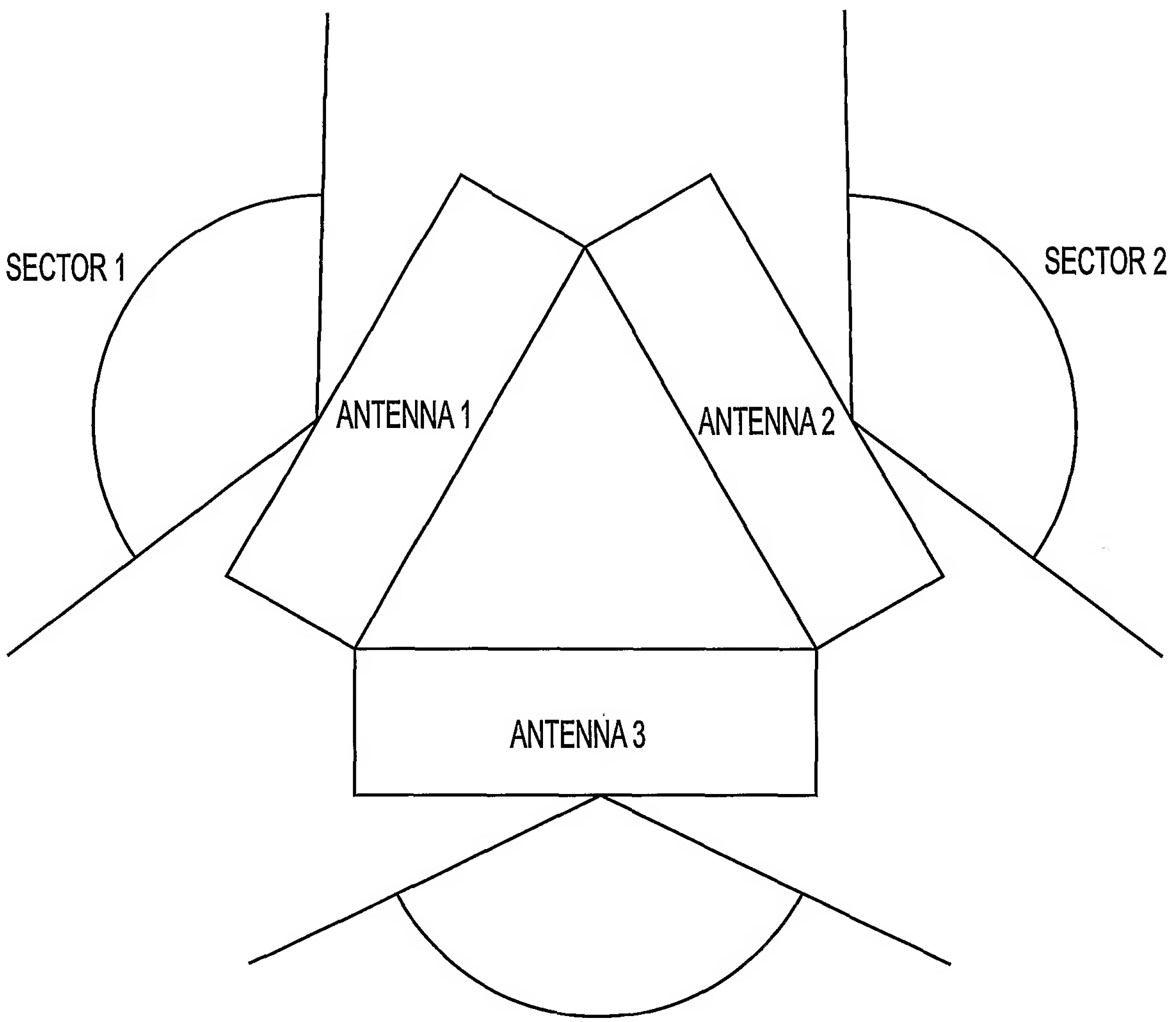


FIG. 8
(PRIOR ART)

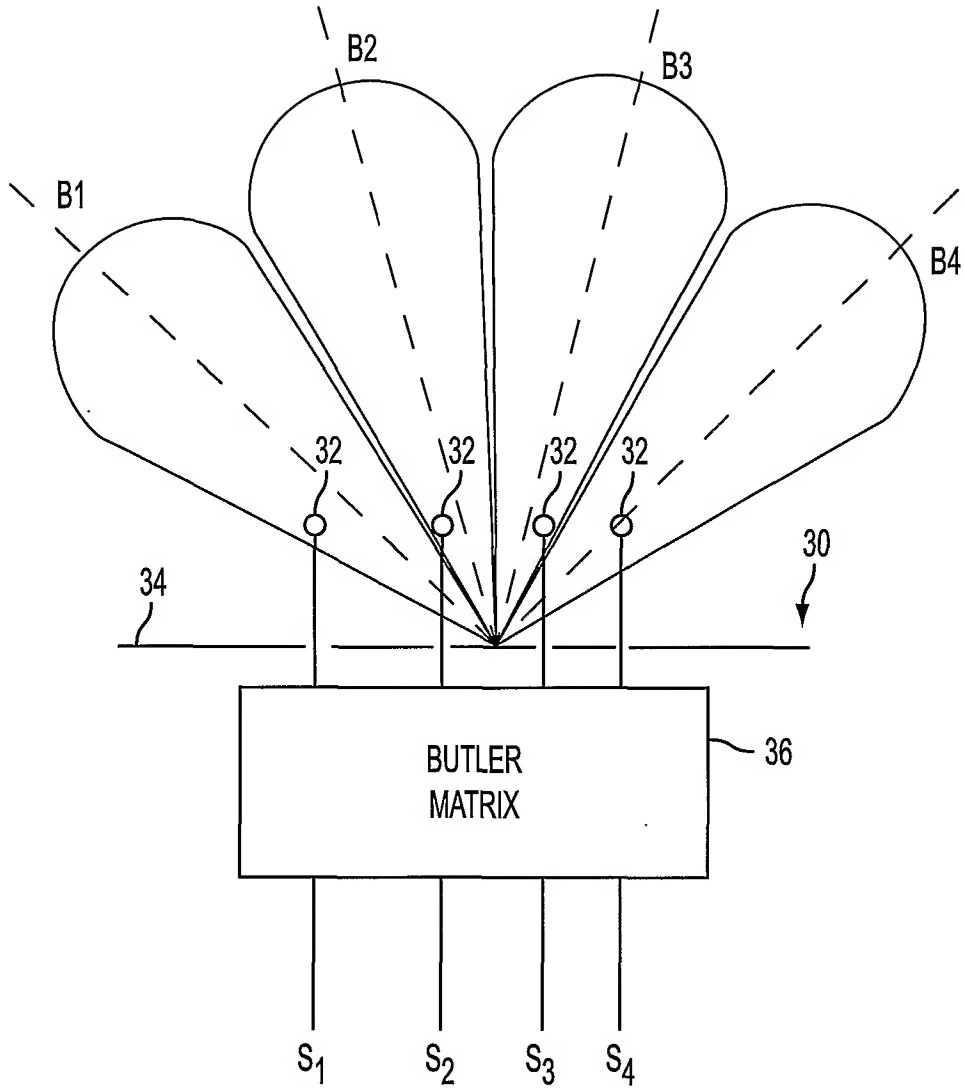


FIG. 9
(PRIOR ART)

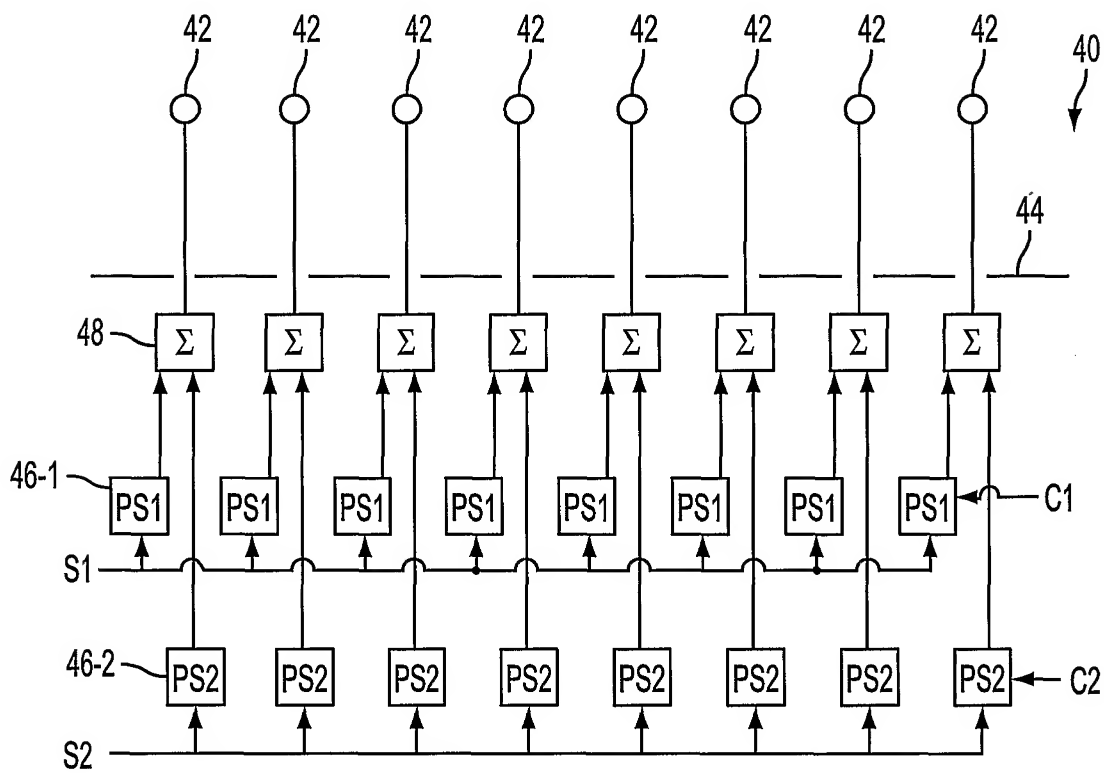


FIG. 10
(PRIOR ART)

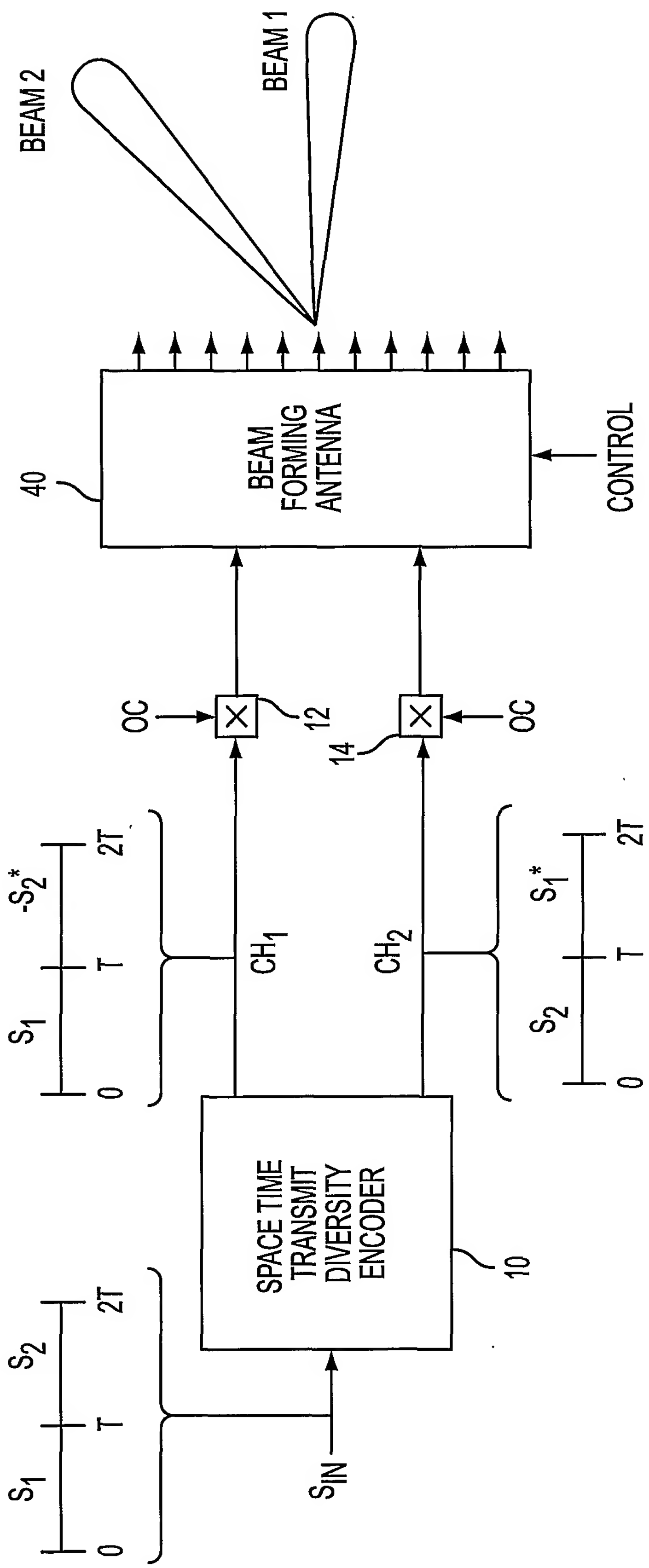

BEAM 2
BEAM 1
40
BEAM FORMING ANTENNA
CONTROL
OC
12
14
OC
S_1
$-S_2^*$
0
T
2T
CH_1
CH_2
S_2
S_1^*
0
T
2T
SPACE TIME TRANSMIT DIVERSITY ENCODER
10
S_1
S_2
0
T
2T
S_{IN}

FIG. 11

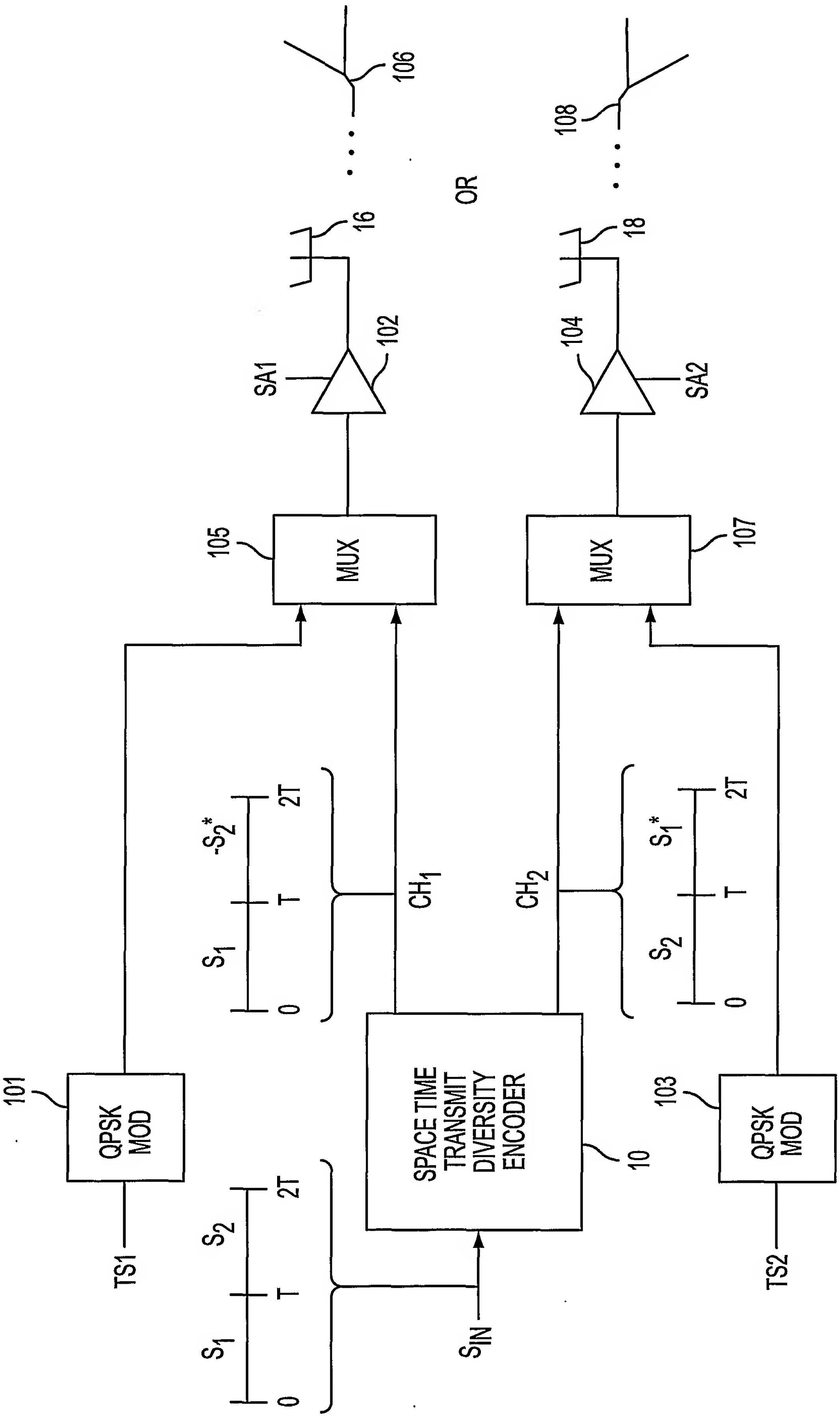

TS1
QPSK MOD
101
S_1
T
S_2
2T
0
S_{IN}
SPACE TIME TRANSMIT DIVERSITY ENCODER
10
TS2
QPSK MOD
103
S_1
$-S_2^*$
CH_1
CH_2
S_2
S_1^*
MUX
105
MUX
107
SA1
102
104
SA2
16
18
106
108
OR

FIG. 12

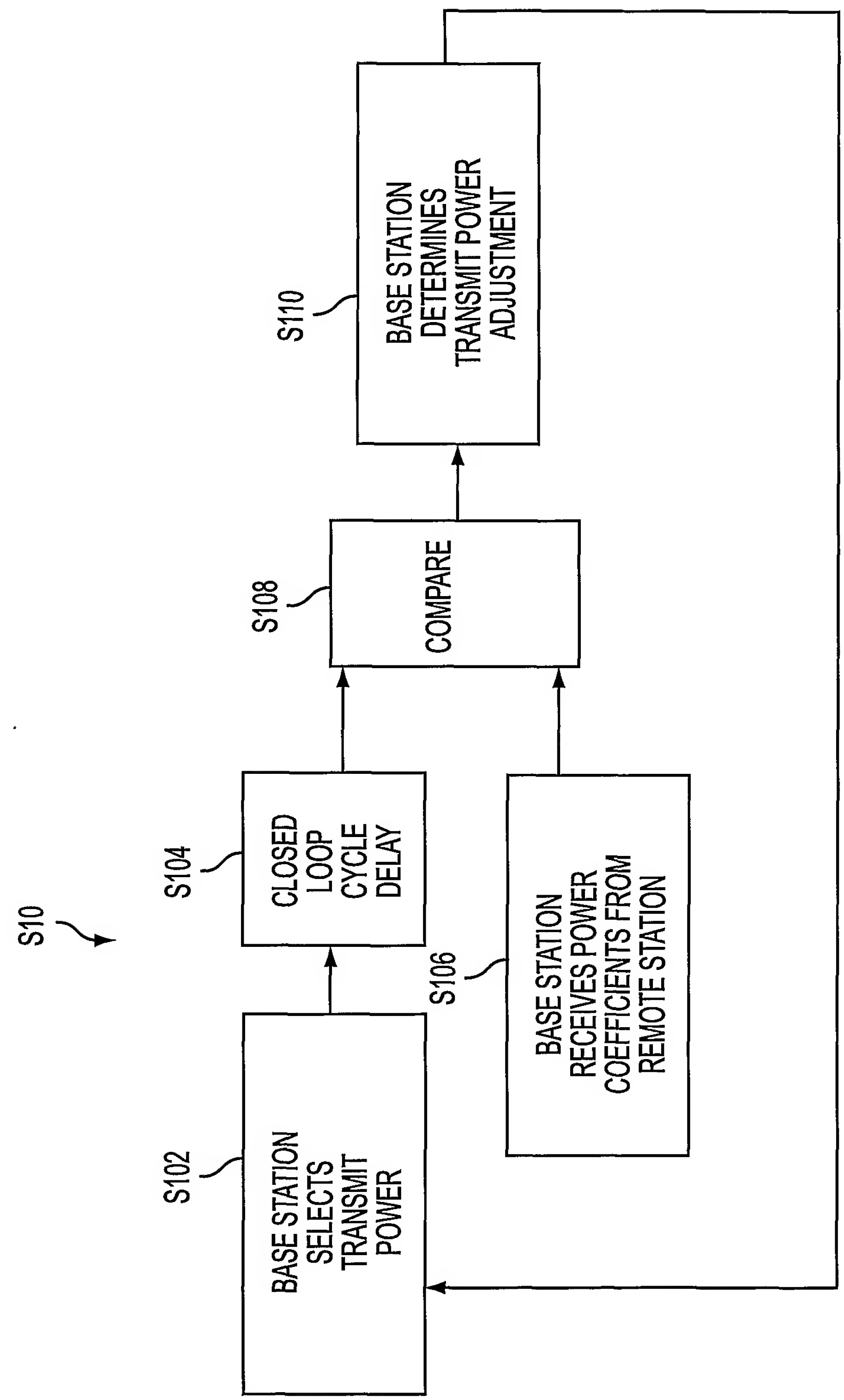

S10
S102
BASE STATION SELECTS TRANSMIT POWER
S104
CLOSED LOOP CYCLE DELAY
S106
BASE STATION RECEIVES POWER COEFFICIENTS FROM REMOTE STATION
S108
COMPARE
S110
BASE STATION DETERMINES TRANSMIT POWER ADJUSTMENT

FIG. 13

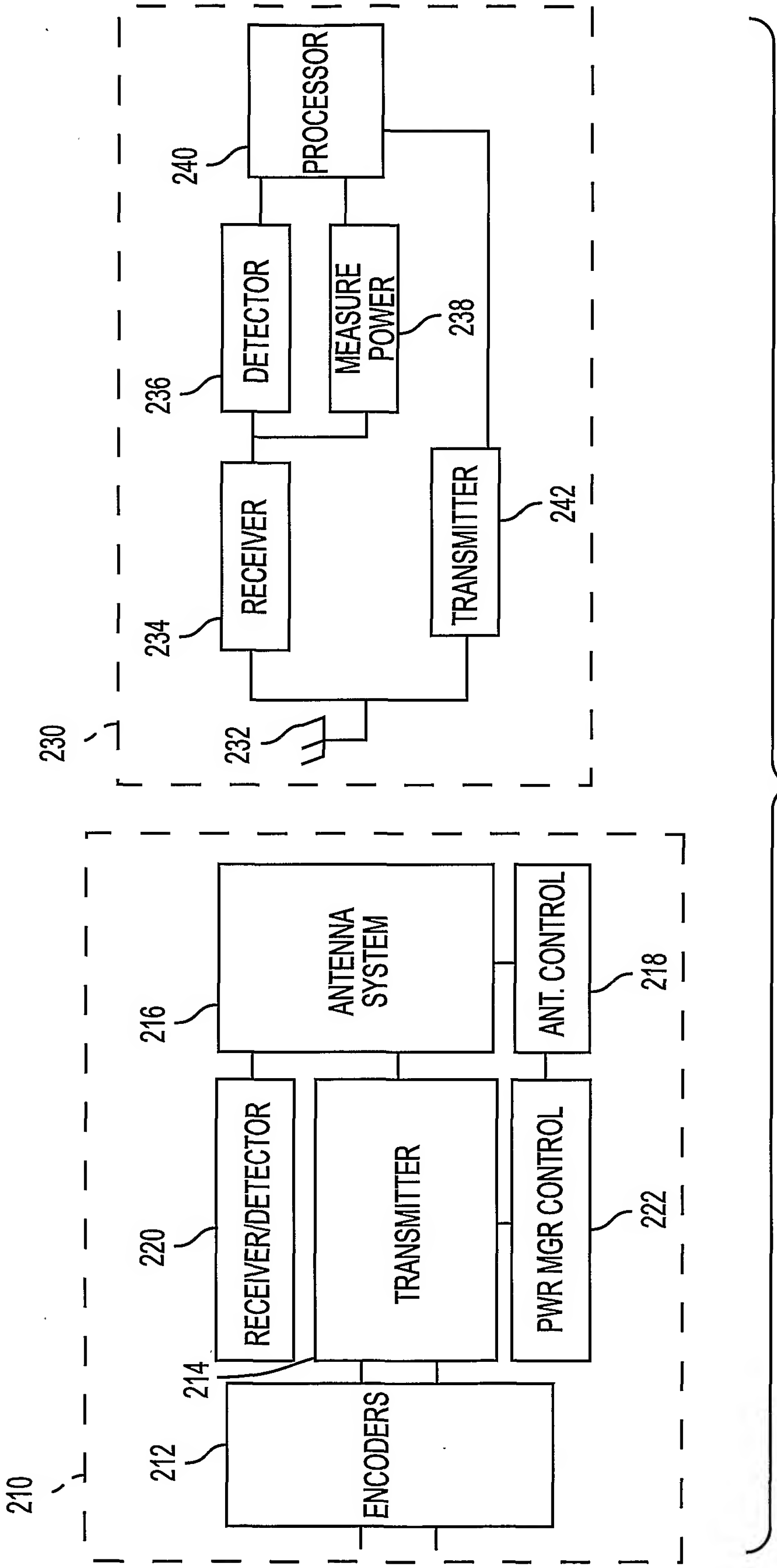

PROCESSOR
240
DETECTOR
236
MEASURE POWER
238
RECEIVER
234
TRANSMITTER
242
230
232
ANTENNA SYSTEM
216
ANT. CONTROL
218
RECEIVER/DETECTOR
220
TRANSMITTER
PWR MGR CONTROL
222
214
ENCODERS
212
210

FIG. 14

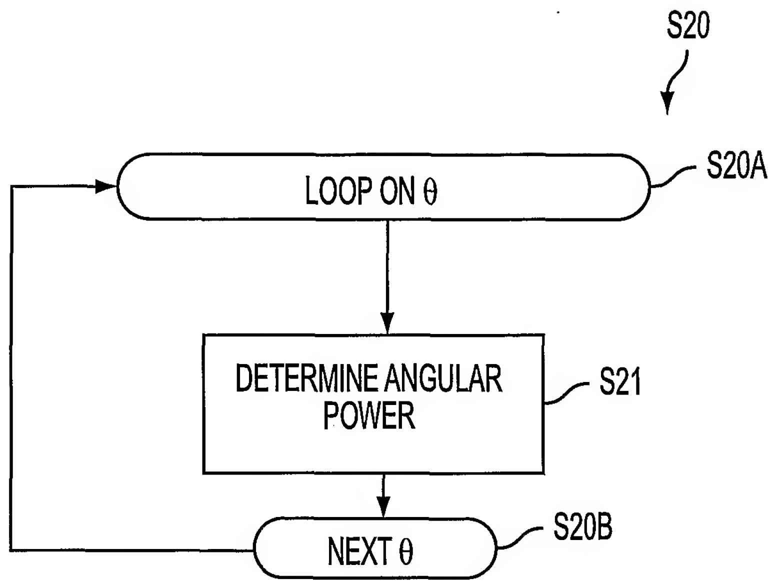

S20
LOOP ON θ
S20A
DETERMINE ANGULAR POWER
S21
NEXT θ
S20B

FIG. 15

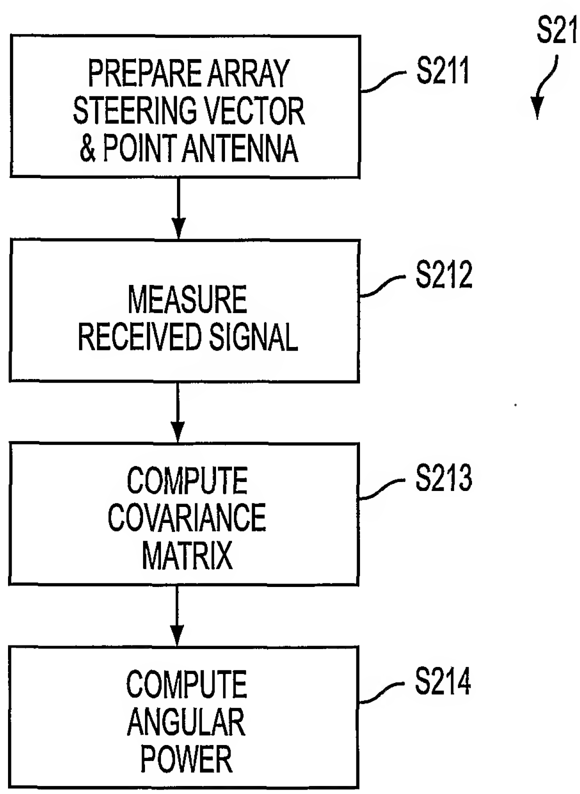

S21
PREPARE ARRAY STEERING VECTOR & POINT ANTENNA
S211
MEASURE RECEIVED SIGNAL
S212
COMPUTE COVARIANCE MATRIX
S213
COMPUTE ANGULAR POWER
S214

FIG. 16

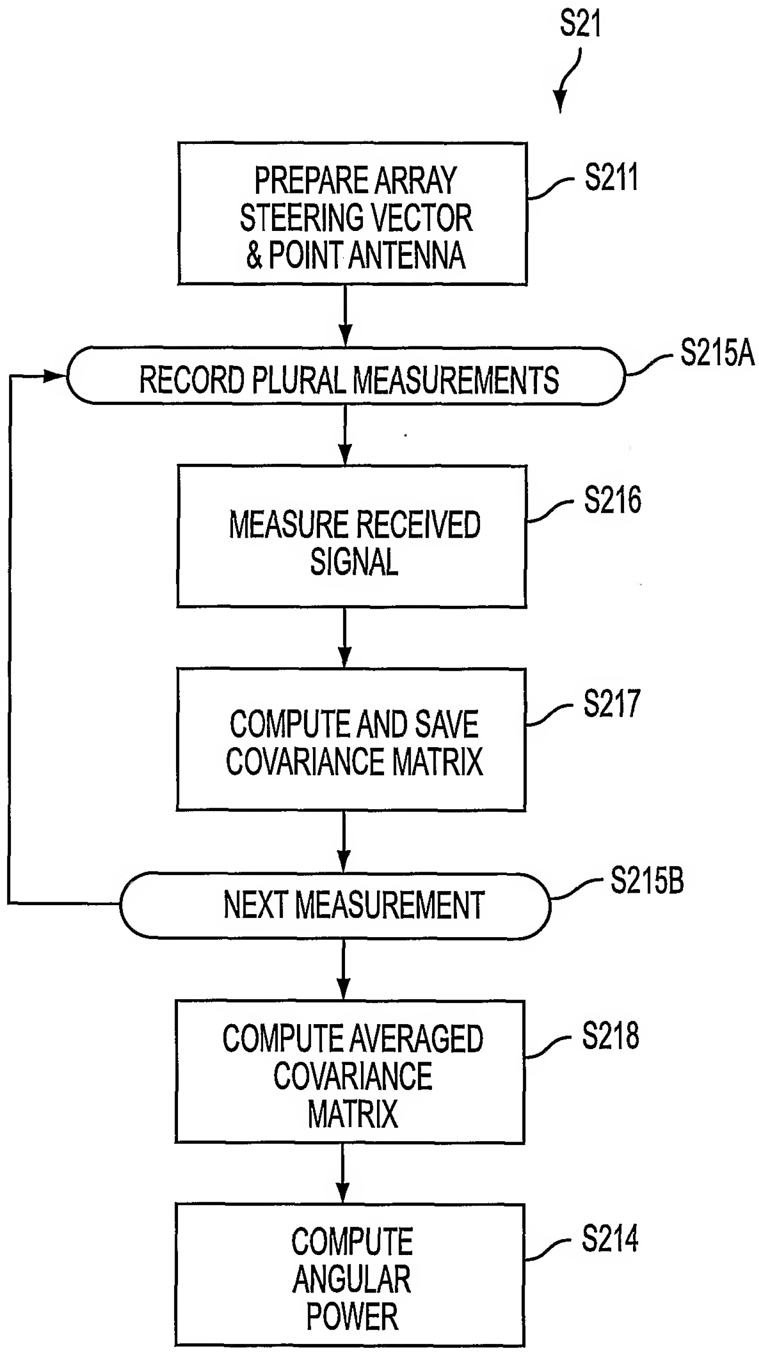

S21
PREPARE ARRAY STEERING VECTOR & POINT ANTENNA
S211
RECORD PLURAL MEASUREMENTS
S215A
MEASURE RECEIVED SIGNAL
S216
COMPUTE AND SAVE COVARIANCE MATRIX
S217
NEXT MEASUREMENT
S215B
COMPUTE AVERAGED COVARIANCE MATRIX
S218
COMPUTE ANGULAR POWER
S214

FIG. 17

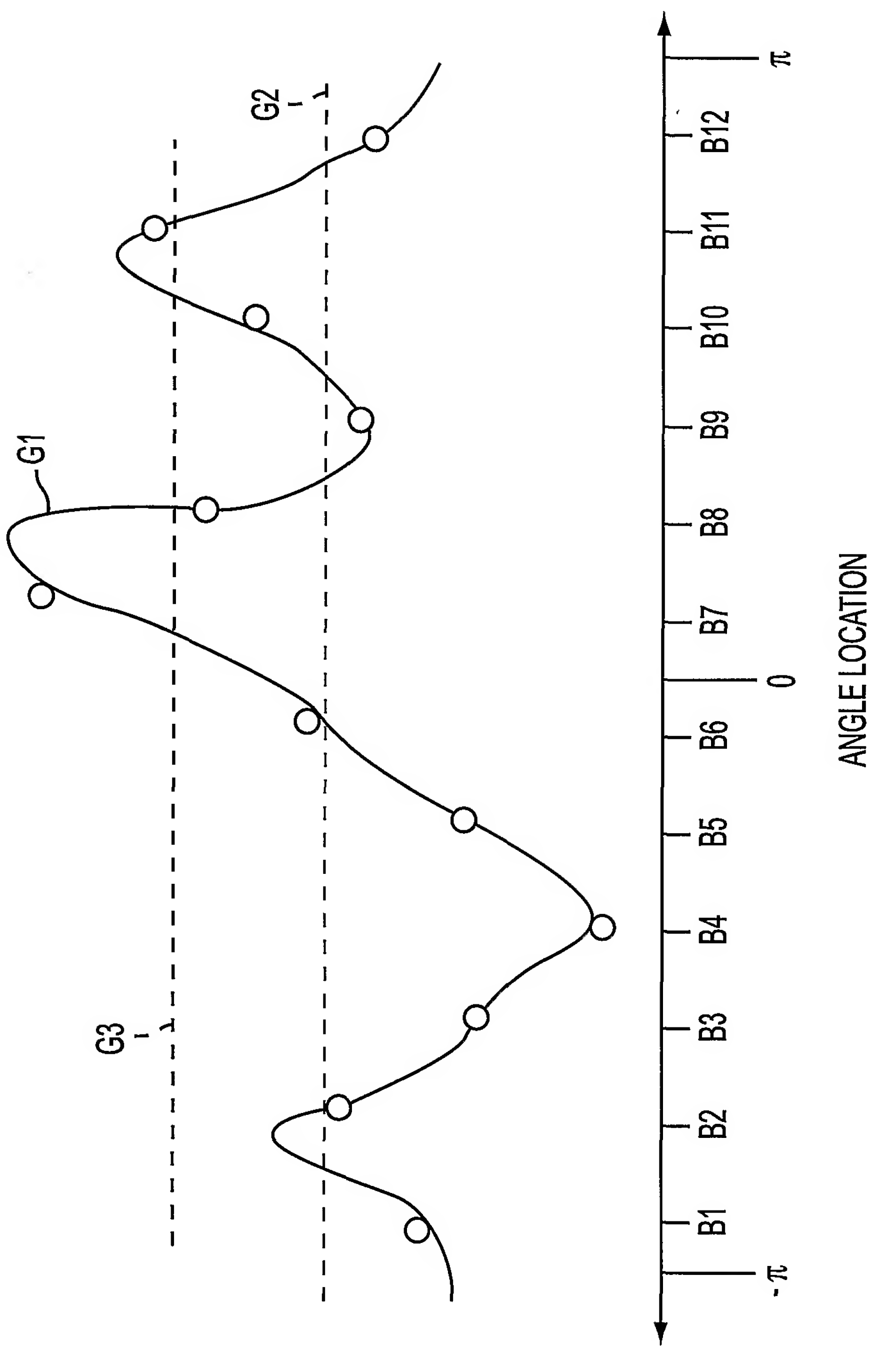


FIG. 18

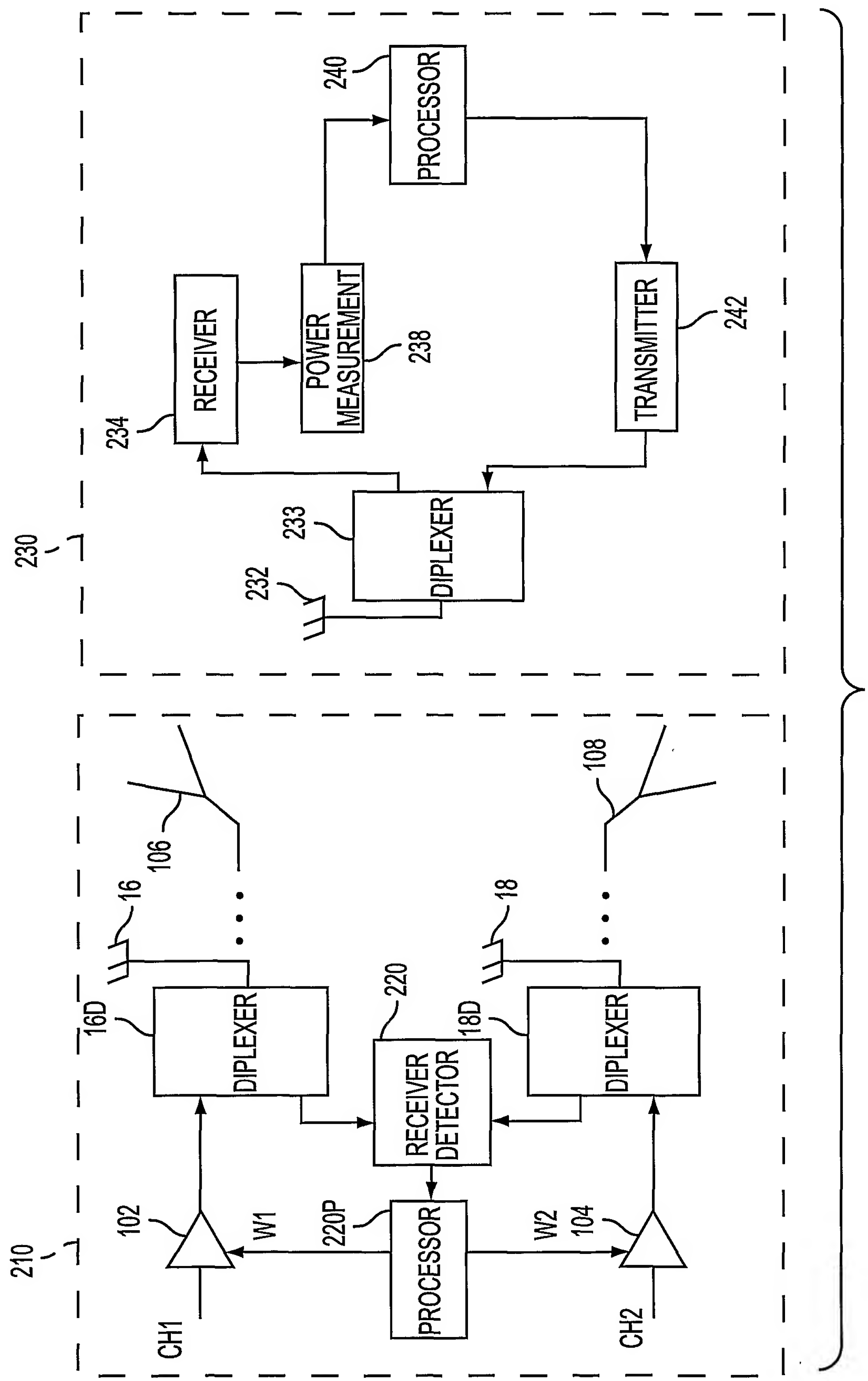

230
232
233
DIPLEXER
234
RECEIVER
POWER MEASUREMENT
238
240
PROCESSOR
TRANSMITTER
242
210
CH1
102
W1
220P
PROCESSOR
W2
104
CH2
RECEIVER DETECTOR
220
16D
DIPLEXER
16
106
18D
DIPLEXER
18
108

FIG. 19

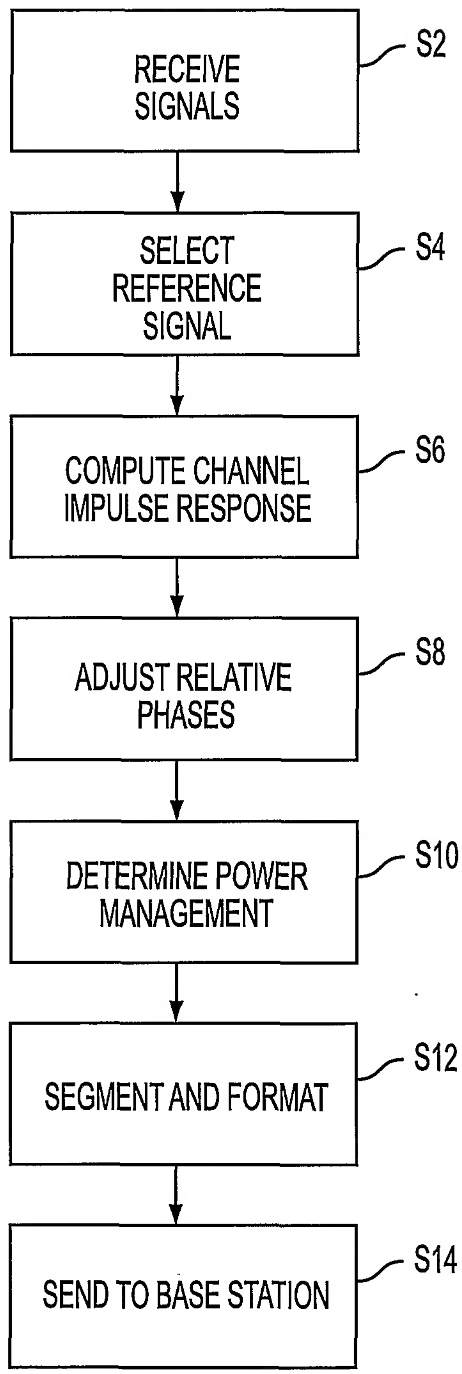

RECEIVE SIGNALS
S2
SELECT REFERENCE SIGNAL
S4
COMPUTE CHANNEL IMPULSE RESPONSE
S6
ADJUST RELATIVE PHASES
S8
DETERMINE POWER MANAGEMENT
S10
SEGMENT AND FORMAT
S12
SEND TO BASE STATION
S14

FIG. 20

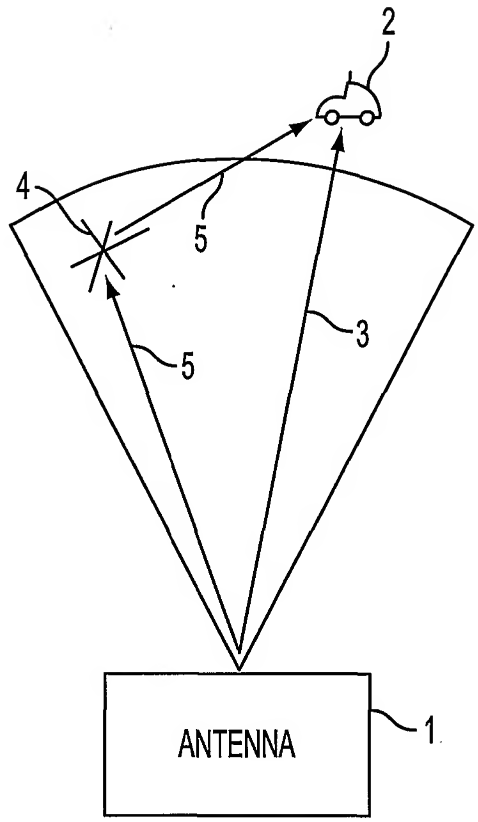


FIG. 21

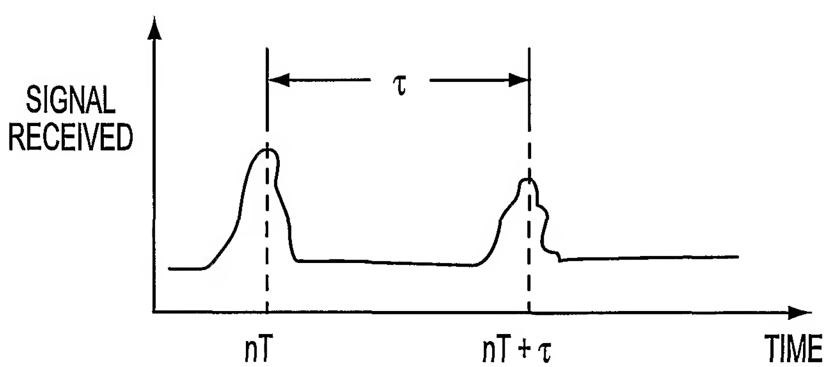


FIG. 22

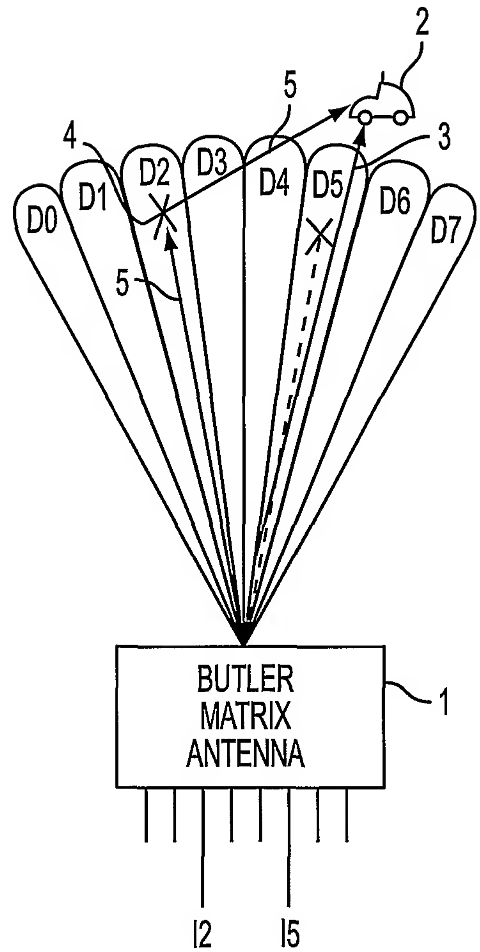


FIG. 23

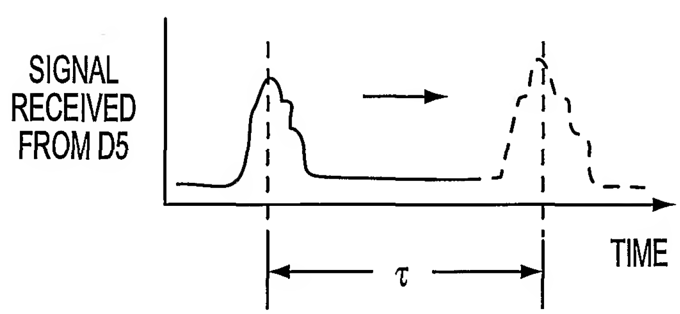


FIG. 24

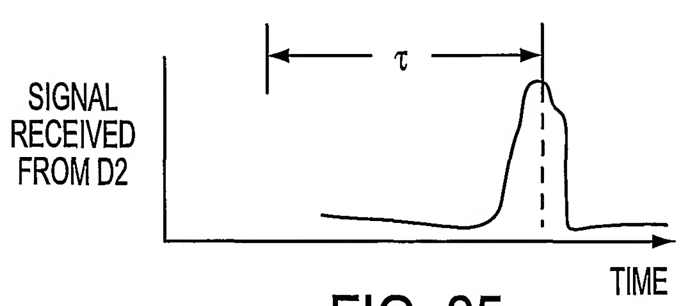


FIG. 25

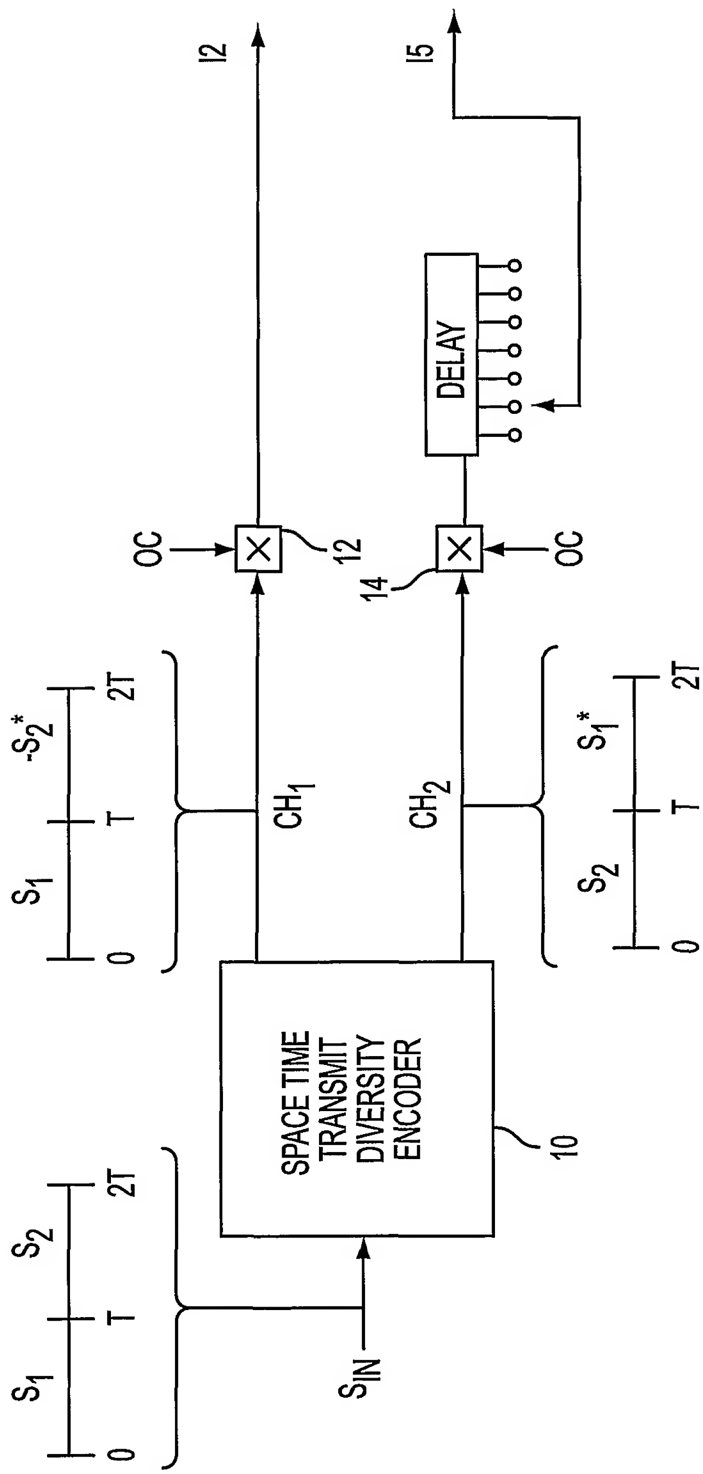

S_1
S_2
0
T
2T
S_{IN}
SPACE TIME TRANSMIT DIVERSITY ENCODER
10
S_1
$-S_2^*$
CH_1
CH_2
S_2
S_1^*
OC
12
14
OC
DELAY
I2
I5

FIG. 26

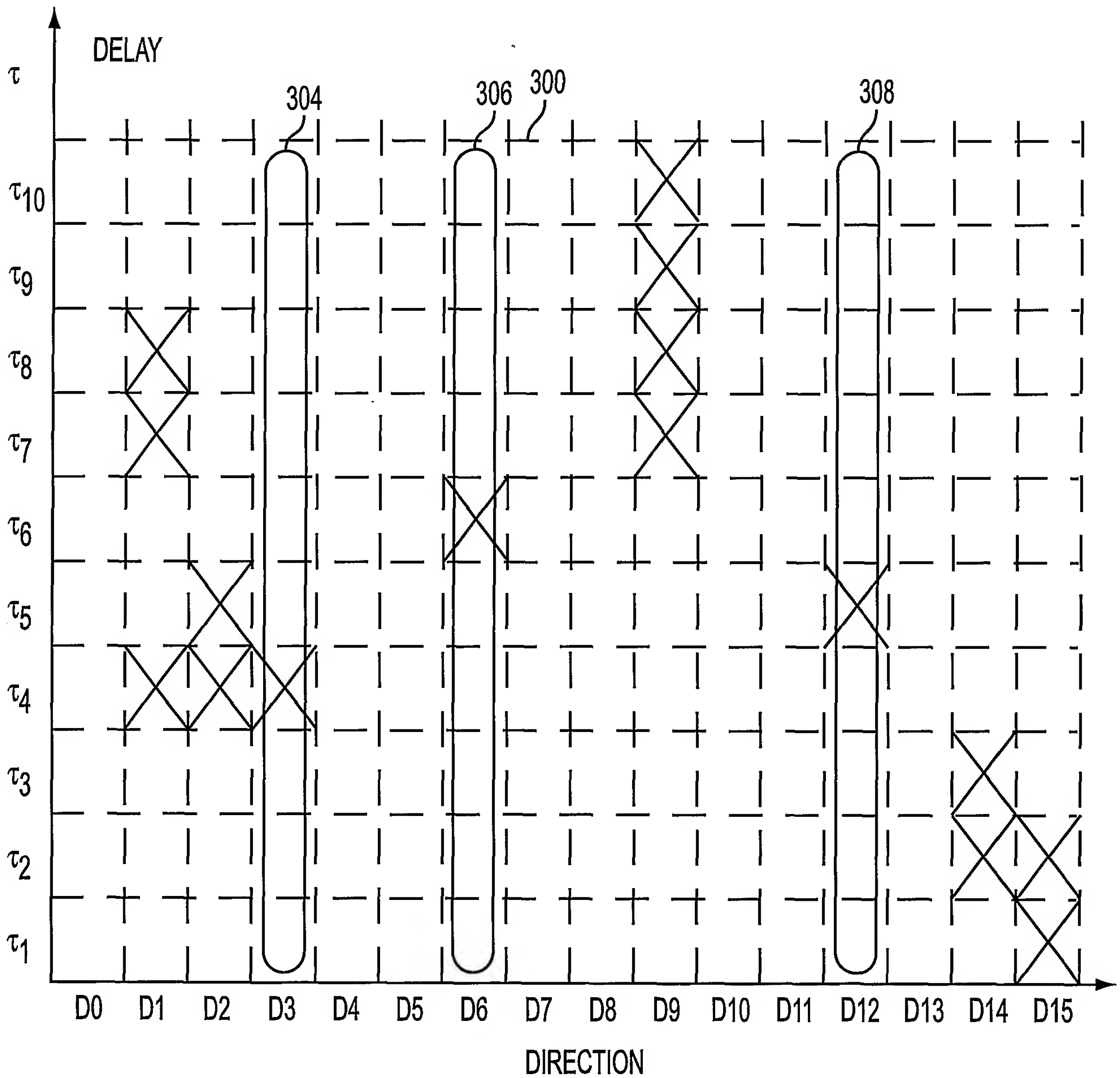

DELAY
τ
τ10
τ9
τ8
τ7
τ6
τ5
τ4
τ3
τ2
τ1
304
306
300
308
D0
D1
D2
D3
D4
D5
D6
D7
D8
D9
D10
D11
D12
D13
D14
D15
DIRECTION

FIG. 27

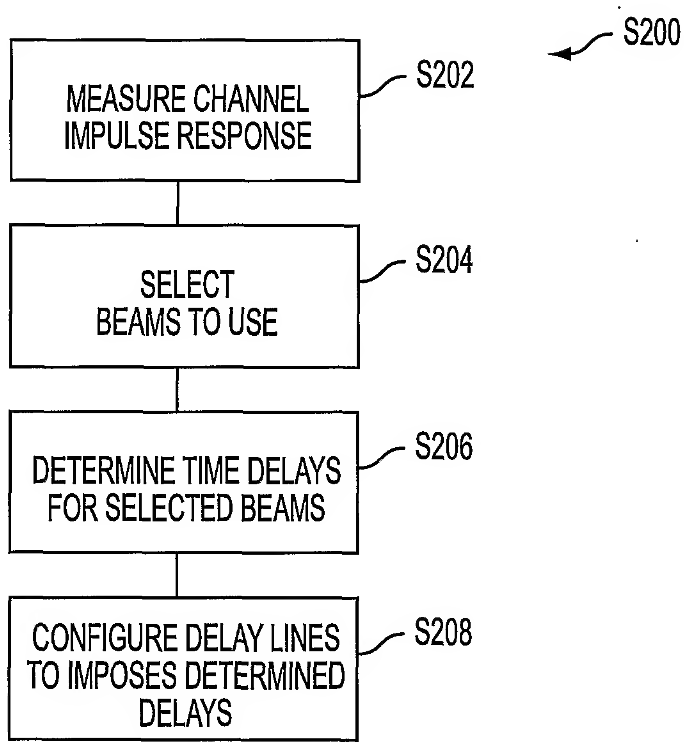

S200
MEASURE CHANNEL IMPULSE RESPONSE
S202
SELECT BEAMS TO USE
S204
DETERMINE TIME DELAYS FOR SELECTED BEAMS
S206
CONFIGURE DELAY LINES TO IMPOSES DETERMINED DELAYS
S208

FIG. 28

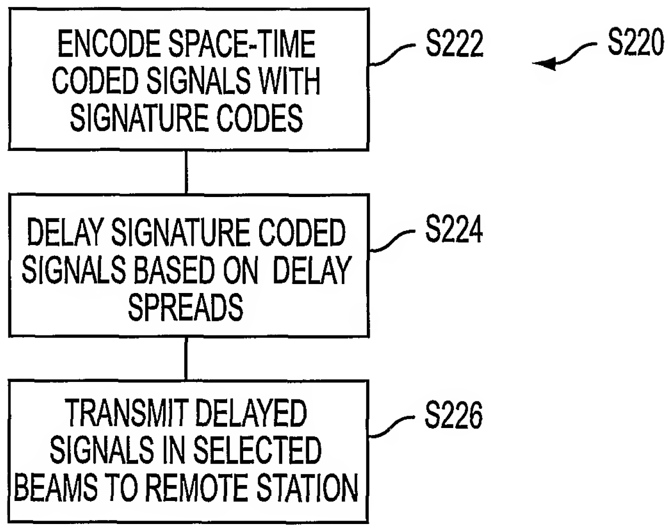

ENCODE SPACE-TIME CODED SIGNALS WITH SIGNATURE CODES
S222
S220
DELAY SIGNATURE CODED SIGNALS BASED ON DELAY SPREADS
S224
TRANSMIT DELAYED SIGNALS IN SELECTED BEAMS TO REMOTE STATION
S226

FIG. 29

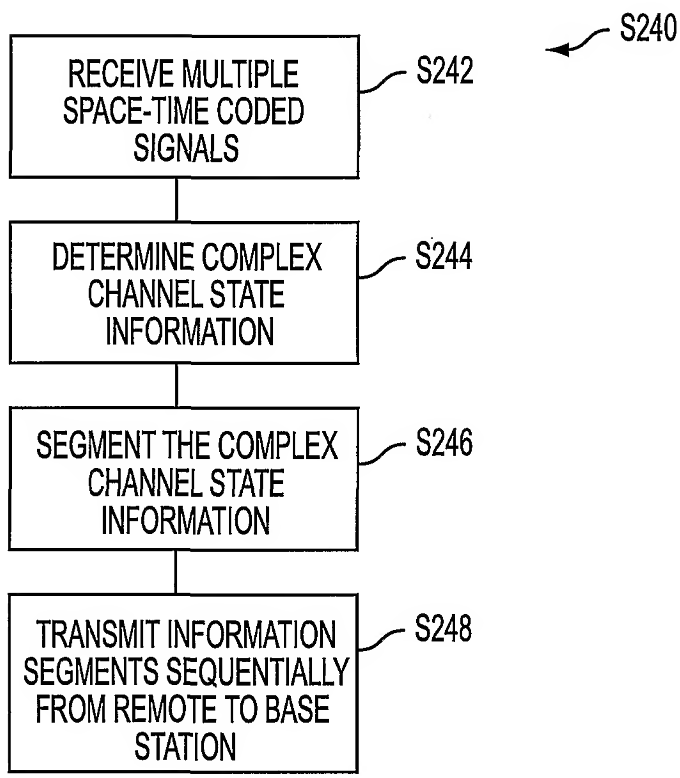

S240
RECEIVE MULTIPLE SPACE-TIME CODED SIGNALS
S242
DETERMINE COMPLEX CHANNEL STATE INFORMATION
S244
SEGMENT THE COMPLEX CHANNEL STATE INFORMATION
S246
TRANSMIT INFORMATION SEGMENTS SEQUENTIALLY FROM REMOTE TO BASE STATION
S248

FIG. 30

INTERNATIONAL SEARCH REPORT

International Application No
PCT/IB 01/00967

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04B7/06 H04B7/04

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, INSPEC, WPI Data

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	R WICHMAN; A HOTTINEN: "Transmit Diversity in WCDMA System" INTERNATIONAL JOURNAL OF WIRELESS INFORMATION NETWORKS, vol. 6, no. 3, 1999, pages 171-180, XP001038053	1-17, 26-42
Y	page 174, paragraphs 3.2,3.2.1 page 175, paragraph 3.3 page 176, paragraph 3.3.2 page 177, paragraph 3.4.1 ---	18,19, 43,44

-/--

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
21 May 2002	21.06.02

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Bodin, C-M

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	JONGREN ET AL: "Combining transmit antenna weights and orthogonal space-time block codes by utilizing side information" SIGNALS, SYSTEMS, AND COMPUTERS, 1999. CONFERENCE RECORD OF THE THIRTY-THIRD ASILOMAR CONFERENCE ON OCT. 24-27, 1999, PISCATAWAY, NJ, USA,IEEE, US, 24 October 1999 (1999-10-24), pages 1562-1566, XP010373893 ISBN: 0-7803-5700-0 page 1562, column 1, line 15 -page 1563, column 2, line 39 ---	1-17, 26-42
X	HEATH ET AL: "Multiple antenna arrays for transmitter diversity and space-time coding" COMMUNICATIONS, 1999. ICC '99. 1999 IEEE INTERNATIONAL CONFERENCE ON VANCOUVER, BC, CANADA 6-10 JUNE 1999, PISCATAWAY, NJ, USA,IEEE, US, 6 June 1999 (1999-06-06), pages 36-40, XP010333801 ISBN: 0-7803-5284-X page 36, column 1, line 1 -page 38, column 1, line 38 ---	1-17, 26-42
A	HOTTINEN ET AL: "Transmit diversity using filtered feedback weights in the FDD/WCDMA system" BROADBAND COMMUNICATIONS, 2000. PROCEEDINGS. 2000 INTERNATIONAL ZURICH SEMINAR ON ZURICH, SWITZERLAND 15-17 FEB. 2000, PISCATAWAY, NJ, USA,IEEE, US, 15 February 2000 (2000-02-15), pages 15-21, XP010376432 ISBN: 0-7803-5977-1 page 16, paragraph 2.2 ---	2-17, 27-42
Y	"3GPP RAN S1.14 V2.0.0, UTRA FDD; physical layer procedures" 3GPP RAN S1.14 V2.0.0, April 1999 (1999-04), pages 1-32, XP002184716 paragraph '08.1! - paragraph '08.4! ---	18,19, 43,44
X	EP 1 003 297 A (LUCENT TECHNOLOGIES INC) 24 May 2000 (2000-05-24) abstract paragraph '0009! - paragraph '0010! column 5, line 36-42 column 8, line 5-9; figures 5,6 ---	20-25, 45-50

-/--

INTERNATIONAL SEARCH REPORT

Int(onal Application No
PCT/IB 01/00967

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	EP 0 969 610 A (LUCENT TECHNOLOGIES INC) 5 January 2000 (2000-01-05) column 3, line 53-57 column 4, line 29-33 column 4, line 45-48 abstract figure 1A ---	20-25, 45-50
X	EP 0 795 970 A (NIPPON ELECTRIC CO) 17 September 1997 (1997-09-17) abstract column 3, line 52-56 figure 1 ---	20-25, 45-50
X	US 5 613 219 A (VOGEL MARTIN DR-ING ET AL) 18 March 1997 (1997-03-18) column 4, line 49 - line 60 -----	20-25, 45-50

INTERNATIONAL SEARCH REPORT

national application No.
PCT/IB 01/00967

Box I Observations where certain claims were found unsearchable (Continuation of item 1 of first sheet)

This International Search Report has not been established in respect of certain claims under Article 17(2)(a) for the following reasons:

1. [ ] Claims Nos.:
because they relate to subject matter not required to be searched by this Authority, namely:

2. [ ] Claims Nos.:
because they relate to parts of the International Application that do not comply with the prescribed requirements to such an extent that no meaningful International Search can be carried out, specifically:

3. [ ] Claims Nos.:
because they are dependent claims and are not drafted in accordance with the second and third sentences of Rule 6.4(a).

Box II Observations where unity of invention is lacking (Continuation of item 2 of first sheet)

This International Searching Authority found multiple inventions in this international application, as follows:

see additional sheet

As a result of the prior review under R. 40.2(e) PCT,
no additional fees are to be refunded.

1. [X] As all required additional search fees were timely paid by the applicant, this International Search Report covers all searchable claims.

2. [ ] As all searchable claims could be searched without effort justifying an additional fee, this Authority did not invite payment of any additional fee.

3. [ ] As only some of the required additional search fees were timely paid by the applicant, this International Search Report covers only those claims for which fees were paid, specifically claims Nos.:

4. [ ] No required additional search fees were timely paid by the applicant. Consequently, this International Search Report is restricted to the invention first mentioned in the claims; it is covered by claims Nos.:

Remark on Protest

[X] The additional search fees were accompanied by the applicant's protest.

[ ] No protest accompanied the payment of additional search fees.

FURTHER INFORMATION CONTINUED FROM PCT/ISA/ 210

This International Searching Authority found multiple (groups of) inventions in this international application, as follows:

1. Claims: 1 - 17 and 26 - 42

   determining complex channel state information based on the received space-time coded signals for feedback to a transmitter

2. Claims: 18, 19, 43 and 44

   transmitting beams with an embedded signature code in a signal from a first station to a second station

   sending an indicia of the selected set of least attenuated signals from the second station to the first station

3. Claims: 20 - 25 and 45 - 50

   determining a time delay associated with a beam at a second station

   setting into a varible delay line the time-delay of the encoder of a first station

INTERNATIONAL SEARCH REPORT

Int(ional Application No
PCT/IB 01/00967

Patent document cited in search report		Publication date	Patent family member(s)			Publication date
EP 1003297	A	24-05-2000	US	6259730	B1	10-07-2001
			BR	9905140	A	10-10-2000
			CN	1253430	A	17-05-2000
			EP	1003297	A2	24-05-2000
			JP	2000151485	A	30-05-2000
			TW	443055	B	23-06-2001
			US	2001019592	A1	06-09-2001
EP 0969610	A	05-01-2000	US	6373832	B1	16-04-2002
			BR	9902616	A	18-01-2000
			EP	0969610	A2	05-01-2000
EP 0795970	A	17-09-1997	JP	2809179	B2	08-10-1998
			JP	9252278	A	22-09-1997
			AU	711362	B2	14-10-1999
			AU	1630097	A	18-09-1997
			CA	2199922	A1	14-09-1997
			EP	0795970	A2	17-09-1997
			US	5905718	A	18-05-1999
US 5613219	A	18-03-1997	DE	4303355	A1	11-08-1994
			EP	0610989	A2	17-08-1994
			JP	6303172	A	28-10-1994
			SG	48179	A1	17-04-1998

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
1 August 2002 (01.08.2002)

PCT

(10) International Publication Number
WO 02/060138 A2

(51) International Patent Classification[7]: H04L 25/02

(21) International Application Number: PCT/US01/50068

(22) International Filing Date: 19 October 2001 (19.10.2001)

(25) Filing Language: English

(26) Publication Language: English

(30) Priority Data:
09/694,432 23 October 2000 (23.10.2000) US

(71) Applicant: **QUALCOMM INCORPORATED** [US/US]; 5775 Morhouse Drive, San Diego, CA 92121-1714 (US).

(72) Inventor: **BLANZ, Josef, J.**; 2315 East Harvard Avenue #7, Denver, CO 80301 (US).

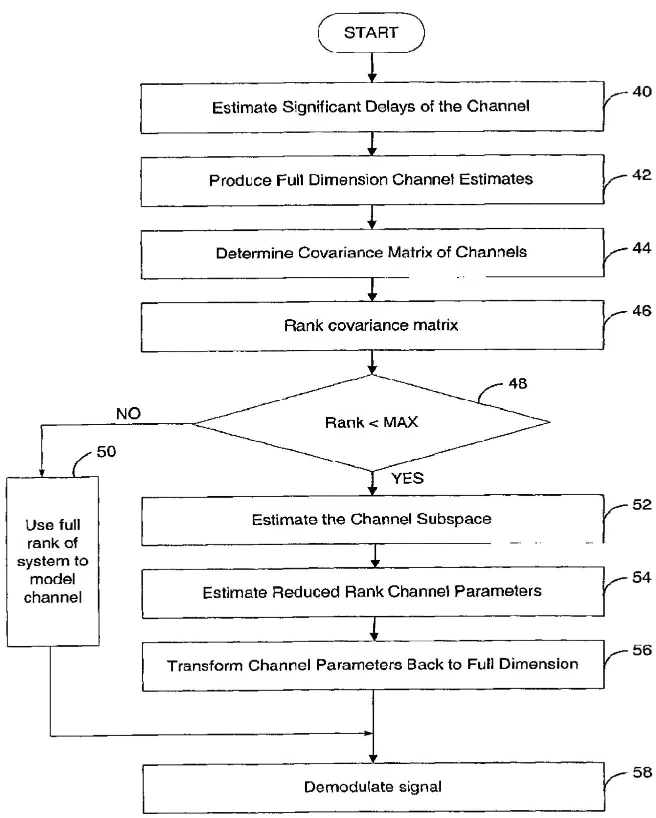
(74) Agents: **WADSWORTH, Philip, R.** et al.; Qualcomm Incorporated, 5775 Morehouse Drive, San Diego, CA 92121-1714 (US).

(81) Designated States *(national)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, PH, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, TZ, UA, UG, UZ, VN, YU, ZA, ZW.

(84) Designated States *(regional)*: ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

[Continued on next page]

(54) Title: METHOD AND APPARATUS FOR REDUCED RANK CHANNEL ESTIMATION IN A COMMUNICATIONS SYSTEM



(57) Abstract: A method and apparatus for estimating a communication channel (14) in a wireless communication system (10) having multiple transmitter antennas (12) using reduced rank estimation. The method exploits redundant and/or a priori knowledge within a system to simplify the estimation calculations. Ine one embodiment, a covariance matrix is calculated (44) and analyzed (46) to determine if the channel parameters maybe reduced for channel estimation. If not, all parameters are used (50), otherwise a reduced rank matrix (54) is used for the calculation.

WO 02/060138 A2

Published:
— *without international search report and to be republished upon receipt of that report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

METHOD AND APPARATUS FOR REDUCED RANK CHANNEL ESTIMATION IN A COMMUNICATIONS SYSTEM

FIELD

The present invention relates to wireless communications. More particularly, the present invention relates to a novel and improved method of reduced rank channel estimation in a communications system.

BACKGROUND

To improve the quality of wireless transmissions, communication systems often employ multiple radiating antenna elements at the transmitter to communicate information to a receiver. The receiver may then have one or more receiver antennas. Multiple antennas are desirable, as wireless communication systems tend to be interference-limited, and the use of multiple antenna elements reduces inter-symbol and co-channel interference introduced during modulation and transmission of radio signals, enhancing the quality of communications. The modeling, and thus design, of such a system, involves estimating several parameters of the space-time channel or link between the transmitter and receiver.

The number of estimated channel parameters per transmit-receiver antenna pair is multiplied by the number of permutations of transmitter-receiver antenna pairs, creating increasingly complicated calculations and decreasing estimation quality. Therefore, it is desirable to have methods of channel estimation that use a reduced set of parameters. Similarly, there is a need for an improved method of channel estimation for radio communications systems having multiple transmitter antennas.

SUMMARY

The presently disclosed embodiments are directed to a novel and improved method and apparatus for estimating channel parameters in a communication link in a wireless communication system having multiple transmitter antennas using a reduced rank estimation method. Each path from a transmitter antenna to the receiver constitutes a channel within the link. The number of channels, therefore, increases with the numbers of transmitter antennas and receiver antennas. The method exploits redundant and/or *a priori* knowledge within a system to simplify the channel model used as a basis for the estimation calculations and to improve the estimation quality. In one embodiment, a covariance matrix is calculated and analyzed to determine if the number of channel parameters may be reduced for channel estimation. If not, all parameters are estimated, otherwise a reduced rank channel model is used for the calculation of channel parameter estimates.

In one aspect, a method for modeling a link in a wireless communication system, the system having a transmitter having N antennas and a receiver having M antennas, each path from one of the N transmitter antennas to the M receiver antennas comprising a channel, includes determining a matrix describing parametric relations of the link; ranking the matrix; determining if the rank is less than NxM; if the rank is less then NxM performing an extraction of a subspace of the matrix; deriving channel impulse responses for each channel based on the extracted subspace of the matrix; and demodulating a received signal using the channel impulse responses. The matrix may be a covariance matrix describing the link, wherein the covariance matrix represents a plurality of impulse responses between the transmitter and the receiver. Alternatively, the matrix may be a sample matrix describing the link.

Further, determining the matrix may include estimating a plurality of parameters describing at least one channel. The parameters may include a distance between transmitter antennas. In one embodiment, the parameters include a transmittal angle with respect to a configuration of the transmitter

antennas. In an alternate embodiment, the matrix describes parametric relations of the link in the frequency domain.

Further, ranking the matrix may include determining an eigenvalue for the matrix. In one embodiment, if the rank is equal to (NxM) a set of correlated impulse responses is applied for demodulating. In one aspect, a wireless apparatus is operative to model a link in a wireless communication system by determining a matrix describing parametric relations of the link; ranking the matrix; determining if the rank is less than NxM; if the rank is less then NxM performing an extraction of a subspace of the matrix; deriving channel impulse responses for each channel based on the extracted subspace of the matrix; and demodulating a received signal using the channel impulse responses.

In another embodiment, a wireless communication apparatus includes a correlator operative to estimate a covariance matrix representing a link with a transmitter based on signals received from the transmitter; a rank analysis unit coupled to the correlator and operative to estimate a rank of the covariance matrix; and a channel estimation unit coupled to the rank analysis unit and operative to generate a reduced rank channel estimate. The covariance matrix may represent a plurality of impulse responses between the apparatus and the transmitter. In one embodiment, the rank analysis unit is operative to determine an eigenvalue corresponding to the covariance matrix and is operative to compare the estimated rank of the covariance matrix to a predetermined full value.

In still another embodiment, a method for estimating a link in a wireless communication system includes estimating a covariance matrix for the link; determining if the rank of the covariance matrix is reducible; reducing the rank of the covariance matrix; and estimating a set of impulse responses for the link using the reduced rank covariance matrix. Additionally, the method may include determining a correlation of the channel; ranking the covariance matrix; and performing an extraction of a reduced rank matrix out of the covariance matrix.

In one embodiment, a wireless communication apparatus is operative within a wireless communication system having a transmitter having N antennas and a receiver having M antennas, each path from one of the N transmitter antennas to the M receiver antennas comprising a channel. The apparatus includes a first set of computer readable instructions operative to determine a covariance matrix describing the link; a second set of computer readable instructions operative to rank the covariance matrix; a third set of computer readable instructions operative to determine if the rank is less than NxM; a fourth set of computer readable instructions operative to perform an extraction of a reduced rank matrix out of the covariance matrix if the rank is less then NxM; a fifth set of computer readable instructions operative to derive channel impulse responses for each channel based on the reduced rank covariance matrix; a sixth set of computer readable instructions operative to demodulate a received signal using the channel impulse responses. The apparatus may further include an equalizer operative in response to the sixth set of computer readable instructions, wherein a configuration of the equalizer is determined by the rank of the covariance matrix. In one embodiment, the apparatus includes a seventh set of computer readable instructions operative to derive a correlated channel impulse response.

In still another aspect, a wireless communication apparatus includes a channel estimation means operative to estimate a covariance matrix representing a link with a transmitter based on signals received from the transmitter; a rank analysis unit coupled to the correlator and operative to estimate the rank of the covariance matrix; and a channel estimation means coupled to the rank analysis unit and operative to generate a reduced rank channel estimate.

Further in another aspect, a wireless communication apparatus includes a correlator operative to estimate a covariance matrix representing a link with a transmitter based on signals received from the transmitter; a rank analysis unit coupled to the correlator and operative to estimate the rank of the covariance

matrix; and a channel estimation means coupled to the rank analysis unit and operative to generate a reduced rank channel estimate.

In yet another aspect, a method for estimating a link in a wireless communication system includes estimating a covariance matrix for the link; determining if the rank of the covariance matrix is reducible; reducing the rank of the covariance matrix; and estimating a set of impulse responses for the link using the reduced rank covariance matrix. The method may further include determining a correlation of the channel; ranking the covariance matrix; and performing an extraction of a reduced rank matrix out of the covariance matrix.

In another embodiment, a wireless apparatus include channel estimation means operative to determine significant delays and determine a set of estimates of full dimension channel parameters associated with the significant delays, wherein each one of the set of estimates corresponds to an instance in time; eigenvalue computation means operative to determine eigenvalues of the set of estimates of the full dimension channel parameters and find any dominant eigenvalues; and channel estimation means operative to determine a set of reduced rank channel parameter estimates in response to the dominant eigenvalues. Further, the apparatus may include eigenvector computation means operative to determine at least one eigenvector associated with one of the dominant eigenvalues of the set of estimates; wherein the channel estimation means uses the at least one eigenvector to project the set of estimates of the full dimension channel parameters onto the subspace spanned by the at least one eigenvector.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, objects, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1 illustrates configurations of wireless communication systems including multiple transmitter antennas;

FIG. 2 illustrates a model of a wireless communication system according to one embodiment;

FIG. 3 illustrates a model of a channel between transmitter and receiver in a wireless communication system;

FIG. 4 illustrates the physical layout of antennas in a transmitter of a wireless communication system;

FIG. 5 illustrates a flow diagram of a method of reduced rank channel estimation for a wireless communication system according to one embodiment;

FIG. 6 illustrates a plot of the estimation gain of one embodiment;

FIG. 7 illustrates a system configuration according to one embodiment.

and

FIG. 8 illustrates an exemplary embodiment of a wireless communication system.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

Multiple radiating antennas may be used to improve transmission quality in a wireless communications system. In the design of third generation mobile radio systems, for example, various transmitter antenna diversity techniques are presented. Multiple transmitter antennas may be used to communicate information to a receiver using a single or multiple receiver antenna(s). Multiple antenna systems offer an improvement in quality. However, the improvement is dependent on the accuracy of the channel model used in the receiver to demodulate the transmitted information. Modeling of the transmission channel uses parameter estimates and determines an effective channel impulse response for the channel. When multiple antennas are used,

the modeling involves estimates of each transmission channel for all transmitter-receiver antenna pairs.

The transmission channel from transmitter to receiver is a space-time channel described generally by at least one impulse response. Often there is little change in the channel parameters from one channel to another, such as where the channel impulse responses differ only in phase. In such a case, it may not be necessary to derive estimates of impulse responses independently for each channel, but rather some information may be reused. When channels are correlated, a reduced rank representation of the channels may be used. Reduced rank refers to the reduced number of completely uncorrelated channels used to describe the link between transmitter and receiver. One way to observe this reduced rank is the rank reduction of the channel covariance matrix used to describe the mutual statistical dependencies of the different channel impulse responses. Note that the reduced rank can also be realized by other parameter measures. For example, in one embodiment a sample matrix is formed of columns comprising samples of channel impulse response estimates over time, wherein the reduced row rank of such a sample matrix is applied as described herein. A reduction in rank may result in a less complex filter or demodulator, *i.e.*, reduces the number of filters and/or filter elements and/or demodulation units used in the receiver. Furthermore, the reduction of the number of estimated parameters used to characterize the channel leads to improved accuracy of the channel model.

FIG. 1 illustrates configurations for wireless communication systems having multiple transmitter Tx antennas. Two paths are illustrated: a first multiple input, multiple output (MIMO) and a second path multiple input single output (MISO). The MISO configuration places multiple Tx antennas in communication with a single Rx antenna. The MIMO configuration extends this to multiple Rx antennas. The channel model for one of the systems of FIG. 1 is illustrated in FIG. 2, in accordance with one embodiment, specifically, for a wireless system employing coherent demodulation, having a link between a transmitter and a receiver, wherein at least the transmitter employs multiple

antennas. The wireless communication system 10 includes a transmitter 12 and receiver 16 that communicate via an air interface. A channel model 14 represents the channels for antenna pairs between transmitter 12 and receiver 16. Channel model 14 considers the channels within a link, such as the MISO link of FIG. 1.

Continuing with FIG. 2, let N_{Tx} be the number of antennas used at the transmitter 12 and N_{Rx} the number at the receiver 16, respectively. In general, for each significant propagation delay between transmitter and receiver, $(N_{Tx} \cdot N_{Rx})$ transmission channels exist for the pair, wherein for a significant propagation delay the received signals resemble the known transmitted signals with high certainty. In other words, define N_E as the number of significant propagation delays, also referred to as echoes. The $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples are then estimated to perform coherent demodulation. When the channels are uncorrelated, the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples are modeled as completely uncorrelated random processes and the estimates of these channel impulse response samples may be derived independently without loss of demodulation performance. However, if the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples are not uncorrelated random processes the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples may be modeled as a linear combination of a smaller number N_{Ch} of channel impulse response samples, wherein $N_{Ch} < (N_{Tx} \cdot N_{Rx} \; N_E)$. Such cases include, but are not limited to, minimal angular spread at the transmitter and/or receiver in the effective channels due to propagation conditions. If N_{Ch} is known, or estimated, and the linear transformation of the N_{Ch} channel impulse response samples are resolved into the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples, then modeling may be accomplished with the N_{Ch} channel impulse response sample estimations. This reduces the number of parameters to be estimated while increasing estimation quality, yielding an increase in the demodulation performance. Even if the exact representation of the linear

transformation of the N_{Ch} channel impulse response samples into the corresponding ($N_{Tx} \cdot N_{Rx} \cdot N_E$) channel impulse response samples is not known, the modeling may still be accomplished with N_{Ch} channel impulse response sample estimations if the subspace spanned by vectors of this linear transformation is known or can be estimated.

This principle is referred to as "Reduced Rank Channel Estimation." The transformation of the N_{Ch} uncorrelated channel impulse responses into the ($N_{Tx} \cdot N_{Rx} \cdot N_E$) correlated channel impulse responses can depend on factors including, but not limited to, the antenna configuration, antenna patterns, polarization characteristics, propagation conditions and more. In some cases the transformation might be known *a priori*, in other cases it can be derived or estimated, for example by angle-of-arrival estimation. The subspace spanned by the linear transformation of the N_{Ch} channel impulse response samples into the corresponding ($N_{Tx} \cdot N_{Rx} \cdot N_E$) channel impulse response samples can be determined by estimating the rank and the eigenvectors of the ($N_{Tx} \cdot N_{Rx} \cdot N_E$)-dimensional covariance matrix of the ($N_{Tx} \cdot N_{Rx} \cdot N_E$) channel impulse response samples. This subspace can also be determined by using a singular value decomposition of a matrix holding columns with all ($N_{Tx} \cdot N_{Rx} \cdot N_E$) channel impulse response sample estimates for different points in time. Note that if the channel impulse response samples are corrupted by a known correlation noise, and if the noise correlation can be estimated, the ($N_{Tx} \cdot N_{Rx} \cdot N_E$) channel impulse response samples may be filtered by a noise de-correlation filter first.

In one embodiment, the rank reducing transformation is known *a priori* or is estimated. In other words, the mapping of the N_{Ch} channels onto the ($N_{Tx} \cdot N_{Rx}$) channels is ascertainable. The reduced rank channel is then estimated using the ascertained transformation. When desired, an equivalent full dimensional channel model may then be derived from the reduced rank estimate by transforming the reduced rank estimate back to the larger dimension.

In an alternate embodiment, the rank reducing transformation is not directly known, but the subspace spanned by the transformation may be extracted from the dominant eigenvectors of the channel covariance matrix. Note that the subspace may be referred to as the signal subspace or the channel subspace. The process involves first estimating a channel covariance matrix and finding the dominant eigenvalues. By determining the associated eigenvectors which span the channel subspace, the process projects the conventional channel estimate into the channel subspace, yielding a reduced rank channel model with reduced estimation errors. If desired, the reduced rank model may be transformed back into an equivalent full dimension channel model.

FIG. 3 illustrates a model 18 of a MIMO channel for continuous time having a linear MIMO filter 20 with N_{Tx} inputs and N_{Rx} outputs. The linear MIMO filter 20 is defined by the $N_{Tx} \times N_{Rx}$ matrix $H(t)$ comprising of linear functions $h_{ij}(t), i = 1...N_{Tx}, j = 1...N_{Rx}$. Generally, $h_{ij}(t), i = 1...N_{Tx}, j = 1...N_{Rx}$ are unknown linear functions. The linear MIMO filter 20 represents the ($N_{Tx} \cdot N_{Rx}$) radio channels through which the N_{Tx} transmit signals pass to the N_{Rx} receiver antennas. These radio channels are characterized by their channel impulse responses $h_{ij}(t), i = 1...N_{Tx}, j = 1...N_{Rx}$. The input signal to the model, $\vec{x}(t)$, is a ($N_{Tx} \times 1$) column vector representing the N_{Tx} band-limited transmit signals, and the output signal from the model, $\vec{y}(t)$, is a ($N_{Rx} \times 1$)column vector , sampled at $t = T, 2T...$, as illustrated by switch T, where the bandwidth of the transmitted signals is less or equal to $1/T$. The received signals contain additive perturbation signals represented by the $N_{Rx} \times 1$ column vector $\vec{z}(t)$, introduced due to noise or co-channel interference. The additive perturbation signals are added at summation nodes 22. The relation between the input signals $\vec{x}(t)$, the channels $H(t)$, the perturbation $\vec{z}(t)$ and the output signals $\vec{y}(t)$ is given by

$$\vec{y}(t) = H^{\mathrm{T}}(t) * \vec{x}(t) + \vec{z}(t), \tag{1}$$

where $*$ denotes the convolution.

FIG. 4 illustrates the physical configuration of antennas at the transmitter of an exemplary embodiment modeled as in FIG. 2. A reduced rank method is applied to estimate the link represented by channel model 14, having a transmitter 12 configured with the four (4) antennas, each spaced at a distance "d." The specifics of the configuration and model are discussed hereinbelow. Note that the estimation procedure is performed at the receiver 16. A reference direction is given by the horizontal line. Angles of transmission are measured with respect to this reference. The angle "α" corresponds to an angle of a propagation path with respect to the reference within a 2-D plane as illustrated. A range of angles with respect to the reference is also illustrated. The following method is used at the receiver 12 in system 10 to estimate the link.

FIG. 5 illustrates a flow diagram of an exemplary method of channel estimation used to process signals in a receiver unit in accordance with one embodiment. Process flow begins by searching for significant propagation delays in the channel, *i.e.* searching for significant echoes at step 40. In one embodiment the process involves a sliding correlation of the received signals with known transmitted signals or known components of the transmitted signals. Correlation refers to the degree with which the received signals are related to the known transmitted signals, wherein a perfect correlation proves a relationship between the signals with high confidence. For time-shifted signals, wherein sliding delays are used to shift the received signals in time, a resultant sliding correlation provides the degree of certainty with which the time-shifted signals resemble the known transmitted signals. Thus in the wireless system context, sliding correlation relates to the synchronization of known signals transmitted by the Tx antennas with time-shifted versions of the received signals. The exemplary embodiment of reduced rank channel estimation uses sliding correlation of the received signals with known transmitted signals to estimate the number N_E and the values $\tau_1, \tau_2, K, \tau_{N_E}$ of significant propagation delays, *i.e.* delays for which the received signals shifted back by these delays in time resemble the known transmitted signals with high certainty. The

procedure of sliding correlation in order to find significant propagation delays is also known as "searching" in CDMA systems.

The method then estimates parameters for multiple observable channels between the N_{Tx} transmitter antennas and the N_{Rx} receiver antennas at step 42. The channels are radio network connection pairs coupling at least a portion of the N_{Tx} transmitter antennas to at least a portion of the N_{Rx} receiver antennas. In the exemplary embodiment, there is a connection between each transmitter 12 antenna and each receiver 16 antenna, resulting in $(N_{Tx} \cdot N_{Rx})$ channels. The parameters describing the multiple channels are those characteristics that impact the impulse responses of the channels. Assuming that N_E significant propagation delays (echoes) exist between transmitter and receiver, $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ complex samples of the $(N_{Tx} \cdot N_{Rx})$ channel impulse responses could be used as a set of parameters describing the multiple channels. This set of parameters is denoted by a $((N_{Tx} \cdot N_{Rx} \cdot N_E)\times 1)$ vector termed $\vec{h}$ herein. The relation between $\vec{x}(t)$, $\vec{h}$, $\vec{z}(t)$, and $\vec{y}(t)$ is developed hereinbelow.

With $\tau_1, \tau_2, \ldots, \tau_{N_E}$ being the significant propagation delays between transmitter and receiver, the model described by (1) can be expressed as

$$\vec{y}(t) = \sum_{e=1}^{N_E} H^{\mathrm{T}}(\tau_e) \cdot \vec{x}(t-\tau_e) + \vec{z}(t). \tag{2}$$

This can be transformed into

$$\vec{y}(t) = \left[\left(\mathrm{I}^{(N_{Rx})} \otimes \vec{x}^{\mathrm{T}}(t-\tau_1)\right)\left(\mathrm{I}^{(N_{Rx})} \otimes \vec{x}^{\mathrm{T}}(t-\tau_2)\right)\ldots\left(\mathrm{I}^{(N_{Rx})} \otimes \vec{x}^{\mathrm{T}}(t-\tau_{N_E})\right)\right] \cdot \vec{h} + \vec{z}(t), \tag{3}$$

where $\otimes$ denotes the Kronecker tensor product, $\mathrm{I}^{(N_{Rx})}$ is a $(N_{Rx} \times N_{Rx})$ identity matrix, and the vector $\vec{h}$ is obtained from the matrix $H(t)$ such that $\vec{h}$

$$\vec{h}_{total}(\tau_e) = vect\{H(\tau_e)\}, \qquad e = 1\ldots N_E, \tag{5}$$

hold. The $((N_{Tx} \cdot N_{Rx})\times 1)$ vector $\vec{h}_{total}(\tau_e)$ is consisting of the elements of the matrix $H(t)$ sampled at τ_e with all columns of $H(\tau_e)$ stacked on top of each

other in the vector $\vec{h}_{total}(\tau_e)$, which is denoted by the operator $vect\{H(\tau_e)\}$ in (5), *i.e.* $\vec{h}_{total}(\tau_e)$ is given by

$$\vec{h}_{total}(\tau_e) = \begin{bmatrix} \left[h_{11}(\tau_e)\ h_{21}(\tau_e) \ldots\ h_{N_{Tx}1}(\tau_e)\right]^{\mathrm{T}} \\ \left[h_{12}(\tau_e)\ h_{22}(\tau_e) \ldots\ h_{N_{Tx}2}(\tau_e)\right]^{\mathrm{T}} \\ \vdots \\ \left[h_{1N_{Rx}}(\tau_e)\ h_{21N_{Rx}}(\tau_e) \ldots\ h_{N_{Tx}N_{Rx}}(\tau_e)\right]^{\mathrm{T}} \end{bmatrix}^{\mathrm{T}}. \qquad (6)$$

Since the output signals $\vec{y}(t)$ are sampled at a sampling rate of $1/T$, vectors containing the discrete time samples can represent segments of a finite duration of the continuous time signals. For the sake of simplicity, the received signals $\vec{y}(t)$ are described herein by a discrete time representation over a finite duration of time $t = 0, T, \ldots, (N_T - 1)T$, where N_T is the number of samples taken over time. Therefore, the following abbreviations are used. Each discrete time transmitted signal at antenna n delayed by τ is given by a vector

$$\vec{s}^{(n)}(\tau) = \left[x_n(0-\tau)\ x_n(T-\tau) \ldots\ x_n((N_T-1)T-\tau)\right]^{\mathrm{T}}, \qquad n = 1 \ldots N_{Tx}. \quad (7)$$

Wherein the matrix describing all of the discrete time transmitted signals delayed by τ is given as

$$S(\tau) = \left[\vec{s}^{(1)}(\tau)\ \vec{s}^{(2)}(\tau) \ldots\ \vec{s}^{(N_{Tx})}(\tau)\right]. \qquad (8)$$

The matrix A describes all discrete time transmit signals having significant delays.

$$A = \left[\left(\mathrm{I}^{(N_{Rx})} \otimes S(\tau_1)\right)\left(\mathrm{I}^{(N_{Rx})} \otimes S(\tau_2)\right) \ldots \left(\mathrm{I}^{(N_{Rx})} \otimes S(\tau_{N_E})\right)\right] \qquad (9)$$

A vector describing each discrete time perturbation signal at antenna n is given as

$$\vec{n}^{(n)} = \left[z_n(0)\ z_n(T) \ldots\ z_n((N_T-1)T)\right]^{\mathrm{T}}, \qquad n = 1 \ldots N_{Rx}. \quad (10)$$

and the vector of all the discrete time perturbation signals is given as

$$\vec{n} = \left[\vec{n}^{(1)\mathrm{T}}\ \vec{n}^{(2)\mathrm{T}} \ldots\ \vec{n}^{(N_{Rx})\mathrm{T}}\right]^{\mathrm{T}}. \qquad (11)$$

The vector of the discrete time received signal at antenna n is given as

$$\vec{r}^{(n)} = \left[y_n(0)\ y_n(T) \ldots\ y_n((N_T-1)T)\right]^{\mathrm{T}}, \qquad n = 1 \ldots N_{Rx}. \quad (12)$$

and the vector of all discrete time received signals is given as

$$\vec{r} = \left[\vec{r}^{(1)\mathrm{T}}\vec{r}^{(2)\mathrm{T}} \ldots \vec{r}^{(N_{Rx})\mathrm{T}}\right]^{\mathrm{T}}. \tag{13}$$

Using the above abbreviations, the discrete time output signals of the MIMO channel model 18 illustrated in FIG. 3 over a period of time from $t = 0, T, \ldots, (N_T - 1)T$ may be reduced to the simple model

$$\vec{r} = A \cdot \vec{h} + \vec{n}. \tag{14}$$

The second step in the flow diagram in FIG. 5 at step 42, is to repeatedly process estimates for a set of parameters characterizing the multiple channels between transmitter and receiver. For the above-described mathematical representation of the channel model, this may be equivalent to processing estimates $\vec{h}^{(n)}$, $n = 1 \ldots N_h$ of the vector $\vec{h}$ in (14) for N_h different points in time. A conventional method uses the correlation of the received signals, shifted back in time by certain delays, with known transmitted signals, such as pilot signals specific to the transmitter antennas, or predetermined training sequences. As the significant propagation delays $\tau_1, \tau_2, \ldots, \tau_{N_E}$ are already determined in step 40, the exemplary embodiment of reduced rank channel estimation uses the correlation of known transmitted signals with versions of the received signals, shifted back in time by $\tau_1, \tau_2, \ldots, \tau_{N_E}$, to generate a channel model, such as channel model 14 of FIG. 2, characterized by the vector $\vec{h}$. If the noise vector $\vec{n}$ represents spatial and temporal white perturbation, wherein the noise covariance matrix is given as $R_n = \langle \vec{n}\vec{n}^{\mathrm{H}} \rangle = \sigma^2 \cdot \mathrm{I}^{(N_{Rx} \cdot N_T)}$, and if the matrix A comprises of the *a priori* known signals, such as pilot symbols of a CDMA system, channel estimates obtained by correlation can be described by

$$\vec{h} = A^{\mathrm{H}} \cdot \vec{r}. \tag{15}$$

If the noise vector $\vec{n}$ does not represent spatial and temporal white perturbation, the channel estimates obtained by correlation can be described by

$$\vec{h} = A^{\mathrm{H}} R_n^{-1} \cdot \vec{r} = A^{\mathrm{H}} R_n^{-1} A \cdot \vec{h} + A^{\mathrm{H}} R_n^{-1} \vec{n}. \tag{16}$$

Note that R_n might be known *a priori* or could be estimated from the received signals. The channel estimate of (16) contains a perturbation vector $A^{\mathrm{H}} R_n^{-1} \vec{n}$

with a covariance matrix of $R_p = A^{\mathrm{H}} R_n^{-1} A$. This covariance matrix is not diagonal in general, *i.e.*, the components of the perturbation vector contained in $\vec{h}$ are correlated in general. If R_p is known or can be estimated, the components of the perturbation vector contained in $\vec{h}$ could be de-correlated by transforming $\vec{h}$ with $R_p^{-1/2}$. This will be assumed in what follows, wherein

$$\vec{h} = R_p^{-1/2} A^{\mathrm{H}} R_n^{-1} \cdot \vec{r} \tag{17}$$

shall hold.

As illustrated in FIG. 5, a covariance matrix of the channel parameters is estimated at step 44. Covariance measures the variance of one random variable with respect to another. In this case, the covariance matrix describes the variance of the various channel parameters with respect to each other. According to the above-described mathematical representation of the channel model, step 44 corresponds to processing an estimate $\hat{R}_h$ of the channel covariance matrix $R_h = \left\langle \vec{h} \cdot \vec{h}^{\mathrm{H}} \right\rangle$. Such an estimate may be given as

$$\hat{R}_h = \frac{1}{N_h} \sum_{n=1}^{N_h} \vec{h}^{(n)} \cdot \vec{h}^{(n)\mathrm{H}} \,. \tag{18}$$

If the MIMO channel has a reduced rank wherein $N_{Ch} < (N_{Tx} \cdot N_{Rx} \cdot N_E)$, *i.e.* the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ MIMO channel impulse response samples can be described as a linear combination of N_{Ch} uncorrelated channel impulse response samples. The channel vector $\vec{h}$ can be modeled as a linear transformation of a channel vector $\vec{g}$ of reduced dimension, wherein

$$\vec{h} = B \cdot \vec{g} \,, \tag{19}$$

and wherein B is a $((N_{Tx} \cdot N_{Rx} \cdot N_E) \times N_{Ch})$ matrix describing the linear transformation. As given hereinabove, the vector $\vec{g}$ is a $(N_{Ch} \times 1)$ vector with uncorrelated components, *i.e.* $R_g = \left\langle \vec{g} \cdot \vec{g}^{\mathrm{H}} \right\rangle$ is a diagonal $(N_{Ch} \times N_{Ch})$ matrix. In this case, the channel covariance matrix is given as

$$R_h = B \cdot R_g \cdot B^{\mathrm{H}} \,. \tag{20}$$

As a consequence, the rank of the channel covariance matrix R_h is equal to N_{Ch}. Given (20), and assuming that correlation according to (17) is used to derive the channel impulse response estimates $\vec{h}$, the covariance matrix of $\vec{h}$ is given as

$$R_{\hat{h}} = \left\langle \vec{h} \cdot \vec{h}^{\mathrm{H}} \right\rangle = R_p^{1/2\mathrm{H}} B \cdot R_g \cdot B^{\mathrm{H}} R_p^{1/2} + \mathrm{I}^{(N_{Tx} \cdot N_{Rx} \cdot N_E)}. \tag{21}$$

Due to the reduced rank N_{Ch} of R_h, the eigenvalue decomposition

$$R_p^{1/2\mathrm{H}} B \cdot R_g \cdot B^{\mathrm{H}} R_p^{1/2} = R_p^{1/2\mathrm{H}} R_h R_p^{1/2} = E \cdot \Lambda \cdot E^{\mathrm{H}}, \tag{22}$$

yields only N_{Ch} non-zero eigenvalues, where Λ is a diagonal matrix containing the eigenvalues and E is a square matrix containing the eigenvectors of $R_p^{1/2\mathrm{H}} B \cdot R_g \cdot B^{\mathrm{H}} R_p^{1/2}$. With (21) and (22) the covariance matrix estimate $\hat{R}_h$ may be expressed by

$$\hat{R}_h = E \cdot (\Lambda + \mathrm{I}^{(N_{Tx} \cdot N_{Rx} \cdot N_E)}) \cdot E^{\mathrm{H}}, \tag{23}$$

i.e., $\hat{R}_h$ shares the eigenvectors with $R_p^{1/2\mathrm{H}} B \cdot R_g \cdot B^{\mathrm{H}} R_p^{1/2}$. Since Λ is a diagonal matrix with only N_{Ch} non-zero elements, $(N_{Tx} \cdot N_{Rx} \cdot N_E) - N_{Ch}$ eigenvalues of $\hat{R}_h$ are constant, and N_{Ch} eigenvalues of $\hat{R}_h$ are larger than the former ones. These larger eigenvalues are termed dominant eigenvalues in the sequel. With a diagonal matrix Λ_C, containing all dominant eigenvalues of the estimated channel covariance matrix, the matrix E_C, containing the corresponding eigenvectors, and with the matrix E_N, containing the remaining eigenvectors, (23) becomes

$$\hat{R}_h = E_C \cdot \Lambda_C \cdot E_C^{\mathrm{H}} + E_N E_N^{\mathrm{H}}. \tag{24}$$

Therefore, the matrix E_C contains the eigenvectors spanning the channel or signal subspace.

The estimated covariance matrix $\hat{R}_h$ is then ranked at step 46, meaning that the number of dominant eigenvalues is estimated. The rank is compared to

a maximum value "MAX" at step 48. MAX is equal to the total number of estimated channel parameters in the vector $\vec{h}$. In other words, MAX is equal to $(N_{Tx} \cdot N_{Rx} \cdot N_E)$. As many of the mechanisms impacting correlation, such as the directionality of the propagation paths, do not change quickly over time, the correlation characteristics may be estimated by averaging over rather long time intervals in comparison to the inverse fading rate of the channel(s).

The rank of the covariance matrix determines whether the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel parameters describing the $(N_{Tx} \cdot N_{Rx})$ existing transmission channels can be modeled as a linear combination of a smaller number N_{Ch} of equivalent uncorrelated channel parameters If a reduced rank is available, the channel subspace E_C of the estimated covariance matrix $\hat{R}_h$ is derived at step 52. Note that instead of using the estimated covariance matrix $\hat{R}_h$, the rank of $\hat{R}_h$ and the channel subspace E_C can also be derived from the matrix of channel parameter estimates

$$\hat{X}_h = \left[\vec{h}^{(1)}\ \vec{h}^{(2)} \ldots\ \vec{h}^{(N_h)}\right], \tag{25}$$

by using singular value decomposition.

With the channel subspace E_C, reduced dimension channel parameter vectors are estimated at step 54, according to

$$\vec{g}^{(n)} = E_C^{\mathrm{H}} \cdot \vec{h}^{(n)}, \tag{26}$$

effectively projecting the originally estimated channel parameters into the channel subspace. This projection into the channel subspace reduces the estimation error. If a reduced complexity demodulator is used in the receiver, which uses the reduced rank channel, *i.e.*, takes only a reduced number of channel parameters into account for demodulation, the estimates of (26) may be directly used in the demodulator for coherent demodulation. In other words, processing would flow directly from step 54 to step 58, or at a minimum step 58 would used the reduced rank estimates.

If a conventional receiver, designed for the full rank channel model, is to be used, the estimates $\hat{g}^{(n)}$ may be transformed back into the full dimensional space at step 56, according to

$$\hat{h}_{new}^{(n)} = R_p^{-1/2H} E_C \cdot \hat{h}^{(n)}, \tag{27}$$

wherein the factor $R_p^{-1/2H}$ is used to make the estimate unbiased. Note that the estimate of the channel subspace E_C may be updated continuously by using a sliding time window for the estimates $\hat{R}_h$ or $\hat{X}_h$, respectively. This eliminates the delay of waiting for a new complete sample set, by using a portion of the previous sample set with incrementally time-shifted new values.

If rank reduction is not possible, processing continues to use the full rank of the system to model the channel at step 50. In this case the method estimates the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel parameters independently from each other. Once the system is modeled, signal demodulation continues at step 58.

The MISO path illustrated in FIG. 1 is provided as an exemplary embodiment. As illustrated, the transmitter, Tx, has four (4) radiating antennas ($N_{Tx} = 4$) and the receiver, Rx, has one (1) antenna ($N_{Rx} = 1$). For simplicity, several assumptions allow a straightforward analysis demonstrating the applicability of the exemplary embodiment to modeling a system as illustrated in FIG. 1. First, the example assumes that each Tx antenna transmits a pilot signal specific to that antenna, wherein the antenna-specific pilot signal is time-aligned and orthogonal to the pilot signals of the other Tx antennas.

Second, assume the channels are frequency non-selective fading channels, each made up of a large number, P, of radio network paths. The paths each have approximately a same run length and a same attenuation. The second assumption ensures that the relative propagation delay is smaller than the inverse of the transmission bandwidth. The propagation delay of two radio paths is typically due to differences in run length.

Third, the channel model is restricted to 2-D propagation, *i.e.* all effective radio paths are located in a 2-D plane. See FIG. 4. Additionally, the geometry of the effective radio paths at the transmitter is assumed to be time-invariant,

wherein each path departure angle, measured with respect to a reference direction of Tx, are concentrated around an average angle, $\overline{\alpha}$. The radio path angles are Gaussian distributed having mean $\overline{\alpha}$ and a standard deviation σ. For one simulation, $\overline{\alpha}$ is selected randomly between –60 and +60 degrees. The standard deviation σ is assumed to be square root of two degrees. Fourth, the arrival paths at Rx are assumed uniformly distributed between 0 and 360 degrees to consider local scattering. Fifth, no line of sight exists.

Sixth, assume a specific phase and Doppler shift for each path. The path-specific phase is selected randomly according to a uniform distribution between 0 and 2π. Additionally, the path-specific phase is adjusted for each Tx antenna according to the geometrical antenna configuration, *i.e.*, the antenna location with respect to a reference point. For phase adjustment, assume object scattering is considered in the far field. The channel-specific Doppler shifts are generated according to a uniform distribution of the angles of arrival paths at Rx, a carrier frequency and a predetermined Rx speed. In the exemplary embodiment, the carrier frequency is assumed to be 1.8 GHz and the receiver speed equal to 60 km/h yielding a maximum Doppler shift of 100 Hz. In the exemplary embodiment, each Tx antenna covers a 120-degree sector, with the antenna patterns all oriented towards $\alpha = 0$.

Given the exemplary system as detailed, application of the process of FIG. 5 provides a channel model having a time variance according to the classic Doppler spectrum. It is possible to consider an antenna-specific radiation pattern. With this channel model, the channel impulse responses for the channels seen through the different transmitter antennas can be generated using the same set of radio paths, thus, introducing realistic correlation in the fading of the different channels.

On the receiving side of the air-interface, at the single antenna of Rx, the method derives an impulse response estimate for each of the four transmission channels, *i.e.*, the four radio network connections between Tx antennas and the Rx antenna. The estimate is based on the *a-priori* knowledge of spreading codes

used to generate the antenna-specific pilot signals associated with each Tx antenna.

Referring again to FIG. 4, in the geographical configuration of antennas at Tx., the antennas are positioned in a line having constant spacing d between neighboring antennas, wherein $d = \lambda$, *i.e.*, antennas are spaced one wavelength apart. Note that Rx has a single omnidirectional antenna. A total number of effective radio paths is considered with $P = 50$. Channel specific variables, α_p, f_p, and Φ_p, represent, respectively, the angle measured from the reference line, the Doppler shift and the phase. The equation describing the channel impulse response for a channel between Tx antenna n and the Rx antenna is given as

$$h_n(t) = \frac{1}{\sqrt{P}} \sum_{p=1}^{P} g_n(\alpha_p) \cdot \exp\left(j \cdot \left(\Phi_p + 2\pi f_p t + \frac{(n - \frac{5}{2})d}{\lambda} \sin(\alpha_p) \right) \right) \cdot \delta(\tau - \tau_0) \qquad (28)$$

wherein $g_n(\alpha)$ is the antenna-specific complex azimuth radiation pattern of each Tx antenna.

If a channel has no angular spread, and all path-specific angles α_p are equal to $\overline{\alpha}$, the channel impulse response for each Tx antenna is given as

$$\begin{aligned} h_n(t) &= g_n(\overline{\alpha}) \cdot \exp\left(j \cdot \frac{(n - \frac{5}{2})d}{\lambda} \sin(\overline{\alpha}) \right) \cdot \frac{1}{\sqrt{P}} \sum_{p=1}^{P} \exp\left(j \cdot (\Phi_p + 2\pi f_p t)\right) \cdot \delta(\tau - \tau_0) \\ &= g_n(\overline{\alpha}) \cdot \exp\left(j \cdot \frac{(n - \frac{5}{2})d}{\lambda} \sin(\overline{\alpha}) \right) \cdot h(t) \end{aligned} \qquad (29)$$

where $h(t)$ is the equivalent channel impulse response for a equivalent isotropic Tx antenna at the reference point.

In this case the channel impulse responses for the different Tx antennas only differ by a complex factor, *i.e.*, the channels are completely correlated. The steering vector is then defined as

$$\vec{a}(\overline{\alpha}) = \begin{pmatrix} g_1(\overline{\alpha}) \cdot \exp\left(j \cdot \frac{-3d}{2\lambda} \sin(\overline{\alpha})\right) \\ g_2(\overline{\alpha}) \cdot \exp\left(j \cdot \frac{-d}{2\lambda} \sin(\overline{\alpha})\right) \\ g_3(\overline{\alpha}) \cdot \exp\left(j \cdot \frac{d}{2\lambda} \sin(\overline{\alpha})\right) \\ g_4(\overline{\alpha}) \cdot \exp\left(j \cdot \frac{3d}{2\lambda} \sin(\overline{\alpha})\right) \end{pmatrix}, \tag{30}$$

and the channel impulse response vector as

$$\vec{h}(t) = \begin{pmatrix} h_1(t) \\ h_2(t) \\ h_3(t) \\ h_4(t) \end{pmatrix}. \tag{31}$$

The four (4) channel impulse responses seen from the Tx antennas are then copies of the channel impulse response $h(t)$, weighted by four different complex factors, which means, the vector $\vec{h}(t)$ is a linear transformation of the scalar $h(t)$ given by

$$\vec{h}(t) = \vec{a}(\overline{\alpha}) \cdot h(t), \tag{32}$$

i.e., the vector $\vec{g}$ in the linear transformation of (19) is in this example equal to the scalar $h(t)$ and the matrix B is equal to the vector $\vec{a}(\overline{\alpha})$. This means the channel covariance matrix $R_h = \langle \vec{h} \cdot \vec{h}^{\mathrm{H}} \rangle$ is equal to $R_h = \vec{a}(\overline{\alpha})\vec{a}(\overline{\alpha})^{\mathrm{H}} \langle |h(t)|^2 \rangle$ in this example. If the steering vector $\vec{a}(\overline{\alpha})$ is known, such as *a-priori* knowledge of the antenna configuration and the radio path direction $\overline{\alpha}$, it is sufficient to estimate the scalar $h(t)$ and either calculate an estimate for $\vec{h}(t)$ using the linear transformation with $\vec{a}(\overline{\alpha})$ or use the estimate of $h(t)$ and $\vec{a}(\overline{\alpha})$ directly for demodulation.

Note that for the case when $\vec{a}(\overline{\alpha})$ is known it may be sufficient to estimate $h(t)$ and then compute an estimate of $\vec{h}(t)$ from the scalar estimate of $h(t)$. If the demodulator is designed such that the channel consists of a single

scalar, *i.e.*, the demodulation considers $\vec{a}(\overline{\alpha})$, then it is possible to demodulate using $\vec{a}(\overline{\alpha})$ and the scalar channel.

The antenna-specific pilot signals at the transmitter are termed $x_n(t)$, $n = 1 \ldots N_{Tx}$, and the relationship is defined by

$$|x_n(t)|^2 \equiv 1 \quad \forall \quad n \in \{1 \ldots N_{Tx}\}. \tag{33}$$

The pilot signals are made up of segments, each having a duration T_S, referred to as the pilot symbol duration, over which the pilot signals are orthogonal, and wherein the following holds

$$\int_{(n-1)T_S}^{nT_S} x_i^*(t) \cdot x_j(t) dt = 0 \quad \forall \quad i, j \in \{1 \ldots N_{Tx}\} \quad i \neq j. \tag{34}$$

The pilot vector is defined as

$$\vec{x}(t) = \begin{pmatrix} x_1(t) \\ x_2(t) \\ x_3(t) \\ x_4(t) \end{pmatrix}, \tag{35}$$

and the receiver noise signal $z(t)$ represents white Gaussian noise. The signal received by the single Rx antenna is described as

$$y(t) = \vec{x}^T(t) \cdot \vec{h}(t) + z(t)., \tag{36}$$

Conventionally, correlating the received signal with the four (4) pilot sequences derives a set of four (4) channel estimates. Wherein the pilot signals are orthogonal over a pilot symbol period, this estimation is then repeated at the pilot symbol rate. Such a correlation procedure is generally referred to as "integrate and dump" and may be expressed as

$$\vec{h}_{\text{cov}}^{(n)} = \frac{1}{T_S} \cdot \int_{(n-1)T_S}^{nT_S} \vec{x}^*(t) \cdot y(t) dt \tag{37}$$

wherein $\vec{h}_{conv}^{(n)}$ is a vector made up of conventional, *i.e.*, integrate and dump, channel estimates derived from the n-th pilot symbol. If (34) is transformed into a discrete time representation, by putting $N_T = T_S / T$ samples of the pilot signals $x_n(t)$ into the columns of the matrix A, N_T samples of the noise signal $z(t)$ into the vector $\vec{n}$, and N_T samples of the received signal $y(t)$ into the vector $\vec{y}$, (34) yields

$$\vec{y} = A \cdot \vec{h}(t) + \vec{n}. \tag{38}$$

Then the discrete time representation of (37) is

$$\vec{h}_{\text{cov}}^{(n)} = \frac{1}{N_T} A^{\mathrm{H}} \cdot \vec{y}. \tag{39}$$

If the channel variations within one pilot symbol are neglected, (39) becomes

$$\vec{h}_{\text{cov}}^{(n)} = \vec{h}(nT_S) + \frac{1}{N_T} \cdot A^{\mathrm{H}} \cdot \vec{n}. \tag{40}$$

Considering the linear transformation of $h(t)$ into $\vec{h}(t)$, the received signal is expressed as

$$\vec{y} = A \cdot \vec{a}(\overline{\alpha}) h(t) + \vec{n}. \tag{41}$$

From this, an estimate of the scalar $h(t)$ is derived as

$$\hat{h}^{(n)} = \frac{\vec{a}^{*T}(\overline{\alpha})}{\left\|\vec{a}(\overline{\alpha})\right\|^2} \cdot \vec{h}_{conv}^{(n)}. \tag{42}$$

Again, when the channel variations within one pilot symbol are ignored, (13) becomes

$$\hat{h}^{(n)} = h(nT_S) + \frac{\vec{a}^{*T}(\overline{\alpha})}{N_T \cdot \left\|\vec{a}(\overline{\alpha})\right\|^2} \cdot A^{\mathrm{H}} \cdot \vec{n}. \tag{43}$$

From this scalar estimate, using the linear transformation, a new estimate of the channel impulse vector is generated as

$$\vec{h}_{new}^{(n)} = \vec{a}(\overline{\alpha})\cdot \hat{h}^{(n)} = \frac{\vec{a}(\overline{\alpha})\cdot \vec{a}^{*T}(\overline{\alpha})}{\left\|\vec{a}(\overline{\alpha})\right\|^2}\cdot \vec{h}_{conv}^{(n)}. \tag{44}$$

Ignoring channel variations within one pilot symbol, (44) becomes

$$\vec{h}_{new}^{(n)} = \vec{h}(nT_S) + \frac{\vec{a}(\overline{\alpha})\cdot \vec{a}^{*T}(\overline{\alpha})}{N_T\cdot\left\|\vec{a}(\overline{\alpha})\right\|^2}\cdot A^{\mathrm{H}}\cdot \vec{n}. \tag{45}$$

If $\vec{a}(\overline{\alpha})$ is not known *a-priori*, it may be estimated using the covariance matrix given by

$$R_h = \left\langle \vec{h}(nT_S)\cdot \vec{h}^{*T}(nT_S)\right\rangle = \vec{a}(\overline{\alpha})\cdot \vec{a}^{*T}(\overline{\alpha})\cdot P_h, \tag{46}$$

with P_h being the average power of the scalar channel impulse response $h(t)$. The covariance matrix R_h can be approximated as

$$\hat{R}_h = \frac{1}{N_{sym}}\sum_{n=1}^{N_{sym}} \vec{h}_{conv}^{(n)}\cdot \vec{h}_{conv}^{(n)\mathrm{H}}, \tag{47}$$

which averages the vector with the conventional channel impulse response estimates over a number, N_{sym}, of pilot symbols.

For the case without noise and having an angular spread equal to zero, $\hat{R}_h$ is rank one (1) and the vector $\vec{a}(\overline{\alpha})$ spans $\hat{R}_h$. Thus (47) reduces to

$$\hat{R}_h\Big|_{noiseless} = R_a = \frac{\vec{a}(\overline{\alpha})\cdot \vec{a}^{*T}(\overline{\alpha})}{\left\|\vec{a}(\overline{\alpha})\right\|^2}. \tag{48}$$

Note that the normalized vector $\vec{a}(\overline{\alpha})/\left\|\vec{a}(\overline{\alpha})\right\|$ spans R_a.

For a noisy case with sufficient low noise power and sufficient low angular spread, $\hat{R}_h$ is still dominated by one eigenvalue. Therefore, the process performs an eigenvalue decomposition of $\hat{R}_h$. When one eigenvalue is much

larger than all other eigenvalues, it is an indication that the angular spread around $\vec{a}(\overline{\alpha})$ was rather small. Therefore, as $\vec{v}_{\max}$ is the eigenvector corresponding to the largest eigenvalue of $\hat{R}_h$, the approximation becomes

$$R_a = \frac{\vec{a}(\overline{\alpha}) \cdot \vec{a}^{*T}(\overline{\alpha})}{\|\vec{a}(\overline{\alpha})\|^2} \approx \vec{v}_{\max} \cdot \vec{v}_{\max}^{*T} \quad . \tag{49}$$

Note that the vector $\vec{v}_{\max}$ in this example is equal to the channel subspace E_C. In general, the estimate $\hat{R}_h$ is used to determine whether the rank of the channel estimation covariance matrix can be reduced. If $\hat{R}_h$ is full rank, the channel estimation problem is not reduced to a smaller dimension.

According to the exemplary embodiment, orthogonal pilot signals of binary chips have a chip rate of 1.2288 Mcps, and a pilot symbol duration of 64 chips. With this channel model, a received signal, including white Gaussian noise, is generated for 4000 consecutive pilot symbols having a pilot Signal-to-Noise Ratio (SNR). From the received signal, 4000 conventional vector estimates, $\vec{h}_{conv}^{(n)}$, are generated. The thus generated covariance matrix $\hat{R}_h$ is averaged over these 4000 consecutive conventional channel estimates. In the exemplary embodiment, the process takes approximately 208.3 ms. After extracting the eigenvector corresponding to the maximal eigenvalue of $\hat{R}_h$, the matrix R_a is calculated. Subsequently, 4000 new vector estimates $\vec{h}_{new}^{(n)}$ are produced according to

$$\vec{h}_{new}^{(n)} = R_a \cdot \vec{h}_{conv}^{(n)} \,. \tag{50}$$

Using the exemplary embodiment, iterations are repeated $N_{\exp} = 50$ times. Over the 50 iterations the transmitter angles are varied such that $\overline{\alpha}$ is uniformly distributed within (+/- 60) degrees, while the angular spread remains constant, having a standard deviation of square root of two ($\sqrt{2}$) degrees. Additionally, the channel parameters for a given pilot SNR are varied.

The varied parameters represent radio path direction(s), path-specific phase, and path-specific Doppler shift, for a certain pilot SNR. An equal number of iterations is performed for different pilot SNR values. A comparison of the quality of the set of conventional estimates to the set of new vector estimates, with respect to the reduction factor of the mean squared estimation error that is averaged over time and iterations, is made using the estimation gain given as

$$g = \frac{\left\langle \left\| \vec{h}_{conv} - \vec{h} \right\|^2 \right\rangle}{\left\langle \left\| \vec{h}_{new} - \vec{h} \right\|^2 \right\rangle}. \tag{51}$$

For the exemplary embodiment, FIG. 6 illustrates the estimation gain in dB as a function of the pilot SNR. Wherein the pilot SNR is defined as the ratio of the average energy per pilot chip E_c of one pilot signal received at the single antenna receiver to the received noise power density I_0 in dB.

The upper limit for the estimation gain is determined by the number of transmit antennas, which is illustrated in FIG. 6 as 6dB. As illustrated in FIG. 6, the estimation gain approaches the upper limit even though the assumed angular spread is not zero and the received signal is severely corrupted by noise. The reduction of the estimation gain with increasing pilot SNR is due to the non-zero angular spread.

Although the channel impulse responses are not completely correlated, the derivation of the impulse response $\vec{h}_{new}$ assumes this property. For larger angular spreads, a smaller estimation gain is expected. For small angular spreads, the estimation gain appears considerable. Note that in general, for residential and suburban environments a standard deviation of one (1) to two (2) degrees is frequently observed. Note also that it is possible to evaluate the performance improvement of the reduced rank channel estimation method using a Monte-Carlo-simulation to derive the reduction of estimation errors of the channel impulse responses as compared to conventional channel estimation using independent correlators.

Reduced rank channel estimation for systems using multiple transmitter antennas allows improvement of the channel estimation quality under certain propagation conditions with limited diversity due to correlated fading. As the mechanisms affecting correlation, such as the directionality of the radio wave propagation, change relatively slowly over time, the correlation characteristics may be estimated by averaging over extended time intervals. This is in contrast to the time intervals associated with inverse fading rate of the channel and thus allows improved accuracy in estimating the correlation characteristics.

Reduced rank channel estimation for multiple transmitter antennas is also applicable to frequency-selective channels by computing either separate estimates of the correlation characteristics or by computing estimates of the correlation characteristics across all propagation delays. Separate estimates refers to computation of $\hat{R}_h$, for each propagation delay. Reduced rank channel estimation is then performed taking into account each delay occurring in the frequency-selective channel impulse response. In an alternate embodiment, wherein additional information, such as the antenna configuration at the transmitter, is known *a-priori*, the step of estimating the linear transformation of the reduced number of uncorrelated channels into the larger number of correlated channels may be more accurate. Additionally, the reduced rank estimation process may be extended to cases with more than one receiver antenna. In this case, the estimation is performed for the MIMO channels, as illustrated in FIG. 1. While the present example involves a system employing coherent demodulation, reduced rank channel estimation as described herein is also applicable to communication systems employing non-coherent demodulation.

A receiver 100 according to one embodiment of the present invention is illustrated in FIG. 7. The receiver 100 has a single antenna 102 that receives signals from a transmitter having multiple antennas. The received signals are first processed by the preprocessor 104. The signals are then provided to a correlator 106, which is used as a sliding correlator for searching and as a correlator for the significant delays for channel estimation. In an alternate

embodiment the delays are determined in software without use of a correlator. The outputs of the correlator 106 are used to provide an estimate of the covariance matrix. In one embodiment, the correlator 106 is made up of fingers to form a rake, having one finger for each combination of transmitter antenna, receiver antenna and significant delay. The estimates are provided to the central processor 112 via bus 116. The processor 112 stores the channel parameter estimates in memory 114 so that the estimates may be used to derive the channel covariance matrix averaged over time.

From memory 114, the estimated covariance matrix is provided to the rank analysis and subspace estimation unit 108 for eigenvalue decomposition. If one or more eigenvalues dominate the others, the channel subspace is estimated by computing the eigenvectors that correspond to the dominant eigenvalues. The eigenvectors spanning the channel subspace are written to memory for further use in the channel subspace projection unit 109 where reduced rank channel parameter estimates are produced by computing the projection of the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ original channel estimates per estimation time interval onto the channel subspace, yielding N_{Ch} reduced rank channel parameter estimates per estimation time interval. The results of the channel subspace projection unit 109 are written to memory for use in the demodulator 110. Optionally the channel subspace projection unit 109 could generate equivalent full dimension channel parameter estimates, by re-transforming the N_{Ch} reduced rank channel parameter estimates into $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ equivalent full dimension channel parameter estimates per estimation time interval. For example in a conventional RAKE-receiver design for the full rank channel model, the number of rake fingers for a full rank demodulator would be $(N_{Tx} \cdot N_{Rx} \cdot N_E)$. A full rank demodulator would then use the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ original channel parameter estimates for the finger coefficients. A reduced complexity demodulator could eventually use only N_{Ch} RAKE fingers using the N_{Ch} reduced rank channel estimates as coefficients. However, since the receiver would generally be designed in anticipation of a worst case situation,

i.e., wherein $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ fingers are implemented, it would be sufficient to compute $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ correlated channel parameter estimates with improved estimation quality over of the N_{Ch} reduced rank channel parameter estimates.

The rank analysis and subspace estimation unit 108 and the subspace projection unit 109 may be implemented in a Digital Signal Processor (DSP), dedicated hardware, software, firmware, or a combination thereof. Modules within receiver 100 may be incorporated together, and are illustrated as separate blocks for clarity based on function.

An exemplary configuration of one embodiment is illustrated in FIG. 8 for a system having four (4) transmitter antennas and two (2) receiver antennas. Three (3) transmission paths are illustrated and labeled 1, 2 and 3. The points of reflection for paths 1 and 2 are both on a same ellipse, wherein the ellipse is formed such that Tx and Rx are the focal points. Note that the ellipse is superimposed on the illustration of the physical layout of the system. Path 3 falls outside of the illustrated ellipse. Paths 1 and 2 have the same significant delay, τ_1, with respect to the receiver, while path 3 has a significant delay, τ_2 different from τ_1. The path delay is a function of the configuration of the antennas as well as the environment of the system. As illustrated, the four (4) transmitter antennas and the two (2) receiver antennas result in eight (8) channels. Each of the path delays, τ_1 and τ_2, produce an echo, wherein $(N_E = 2)$. The dimension of the covariance matrix is given as $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ or sixteen (16) corresponding to the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples. Therefore, the full rank channel parameter vector is a 16-dimension vector. Using the rank reduction methods described herein, the rank of the channel estimation may be reduced to three (3) dimensions, corresponding to paths 1, 2, and 3, wherein $(N_{Ch} = 2)$. Note that where the mapping of the N_{Ch} transmission paths to the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples is not known, the subspace may be extracted from configuration information. If the location and characteristics, such as direction and directionality, of the

antennas are known, the information may be used to generate an array response or steering vector. Using the steering vector and path direction information, which is also extractable using subspace algorithms, the angle of transmission, α, is estimated. If the antenna configuration has a fixed deployment the angle of transmission is calculable. A vector is formed including an angle of transmission for each transmitter antenna. Similarly, an arrival angle vector is formed considering the receiver antennas. A linear transformation for the mapping of the N_{Ch} transmission paths to the $(N_{Tx} \cdot N_{Rx} \cdot N_E)$ channel impulse response samples is constructed using this information from both the transmitter and receiver configurations. This provides the matrix B as given in (19) hereinabove describing the linear transformation. The covariance matrix is derived therefrom as in (20) hereinabove. The process then proceeds as for the case where corresponding information is obtained from *a priori* knowledge.

While one embodiment has been described herein with respect to the time domain, an alternate embodiment performs a rank reduction of the covariance matrix or a sample matrix in the frequency domain. If the parameters and equations are developed in the frequency domain, the process to estimate the channel then incorporates the frequency domain values.

The previous description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. The various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

CLAIMS

I (WE) CLAIM:

1. A method for modeling a link in a wireless communication system, the system having a transmitter having N antennas and a receiver having M antennas, each path from one of the N transmitter antennas to the M receiver antennas comprising a channel, the method comprising:

    determining a matrix describing parametric relations of the link;

    ranking the matrix;

    determining if the rank is less than NxM;

    if the rank is less then NxM performing an extraction of a subspace of the matrix;

    deriving channel impulse responses for each channel based on the extracted subspace of the matrix; and

    demodulating a received signal using the channel impulse responses.

2. The method of claim 1, wherein the matrix is a covariance matrix describing the link, wherein the covariance matrix represents a plurality of impulse responses between the transmitter and the receiver.

3. The method of claim 1, wherein the matrix is a sample matrix describing the link.

4. The method of claim 1, wherein the step of determining the matrix further comprises:

    estimating a plurality of parameters describing at least one channel.

5. The method of claim 4, wherein the parameters include a distance between transmitter antennas.

6. The method of claim 4, wherein the parameters include a transmittal angle with respect to a configuration of the transmitter antennas.

7. The method of claim 4, wherein the determining the matrix comprises estimating the matrix.

8. The method of claim 1, wherein the matrix describes parametric relations of the link in the frequency domain.

9. The method of claim 1, wherein the ranking the matrix, further comprises:

determining an eigenvalue for the matrix.

10. The method of claim 1, wherein if the rank is equal to (NxM) a set of correlated impulse responses is applied for demodulating.

11. A wireless apparatus operative to perform the method of claim 1.

12. A wireless communication apparatus, comprising:

a correlator operative to estimate a covariance matrix representing a link with a transmitter based on signals received from the transmitter;

a rank analysis unit coupled to the correlator and operative to estimate a rank of the covariance matrix; and

a channel estimation unit coupled to the rank analysis unit and operative to generate a reduced rank channel estimate.

13. The apparatus of claim 12, wherein the covariance matrix represents a plurality of impulse responses between the apparatus and the transmitter.

14. The apparatus of claim 12, wherein correlator is operative to determine a correlation of at least two channels.

15. The apparatus of claim 14, wherein the rank analysis unit is operative to determine an eigenvalue corresponding to the covariance matrix.

16. The apparatus of claim 15, wherein the rank analysis unit is operative to compare the estimated rank of the covariance matrix to a predetermined full value.

17. A method for estimating a link in a wireless communication system, the method comprising:

estimating a covariance matrix for the link;

determining if the rank of the covariance matrix is reducible;

reducing the rank of the covariance matrix; and

estimating a set of impulse responses for the link using the reduced rank covariance matrix.

18. The method of claim 17, further comprising:

determining a correlation of the channel;

ranking the covariance matrix; and

performing an extraction of a reduced rank matrix out of the covariance matrix.

19. A wireless communication apparatus operative within a wireless communication system having a transmitter having N antennas and a receiver having M antennas, each path from one of the N transmitter antennas to the M receiver antennas comprising a channel, the apparatus comprising:

a first set of computer readable instructions operative to determine a covariance matrix describing the link;

a second set of computer readable instructions operative to rank the covariance matrix;

a third set of computer readable instructions operative to determine if the rank is less than NxM;

a fourth set of computer readable instructions operative to perform an extraction of a reduced rank matrix out of the covariance matrix if the rank is less then NxM;

a fifth set of computer readable instructions operative to derive channel impulse responses for each channel based on the reduced rank covariance matrix;

a sixth set of computer readable instructions operative to demodulate a received signal using the channel impulse responses.

20. The apparatus of claim 19, further comprising:

an equalizer operative in response to the sixth set of computer readable instructions, wherein a configuration of the equalizer is determined by the rank of the covariance matrix.

21. The apparatus of claim 19, further comprising:

a seventh set of computer readable instructions operative to derive a correlated channel impulse response.

22. A wireless communication apparatus, comprising:

a channel estimation means operative to estimate a covariance matrix representing a link with a transmitter based on signals received from the transmitter;

a rank analysis unit coupled to the correlator and operative to estimate the rank of the covariance matrix; and

a channel estimation means coupled to the rank analysis unit and operative to generate a reduced rank channel estimate.

23. A wireless communication apparatus, comprising:

a correlator operative to estimate a covariance matrix representing a link with a transmitter based on signals received from the transmitter;

a rank analysis unit coupled to the correlator and operative to estimate the rank of the covariance matrix; and

a channel estimation means coupled to the rank analysis unit and operative to generate a reduced rank channel estimate.

24. A method for estimating a link in a wireless communication system, the method comprising:

estimating a covariance matrix for the link;

determining if the rank of the covariance matrix is reducible;

reducing the rank of the covariance matrix; and

estimating a set of impulse responses for the link using the reduced rank covariance matrix.

25. The method of claim 24, further comprising:

determining a correlation of the channel;

ranking the covariance matrix; and

performing an extraction of a reduced rank matrix out of the covariance matrix;

26. A wireless apparatus, comprising:

channel estimation means operative to determine significant delays and determine a set of estimates of full dimension channel parameters associated with the significant delays, wherein each one of the set of estimates corresponds to an instance in time;

eigenvalue computation means operative to determine eigenvalues of the set of estimates of the full dimension channel parameters and find any dominant eigenvalues; and

channel estimation means operative to determine a set of reduced rank channel parameter estimates in response to the dominant eigenvalues.

27. The wireless apparatus of claim 26, further comprising:

eigenvector computation means operative to determine at least one eigenvector associated with one of the dominant eigenvalues of the set of estimates;

wherein the channel estimation means uses the at least one eigenvector to project the set of estimates of the full dimension channel parameters onto the subspace spanned by the at least one eigenvector.

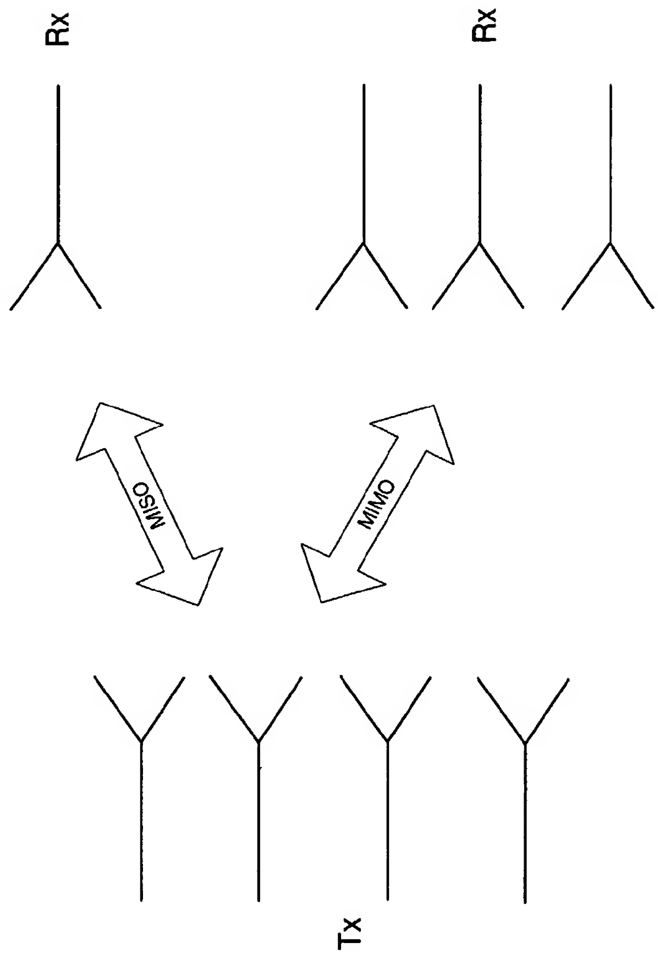


FIG. 1

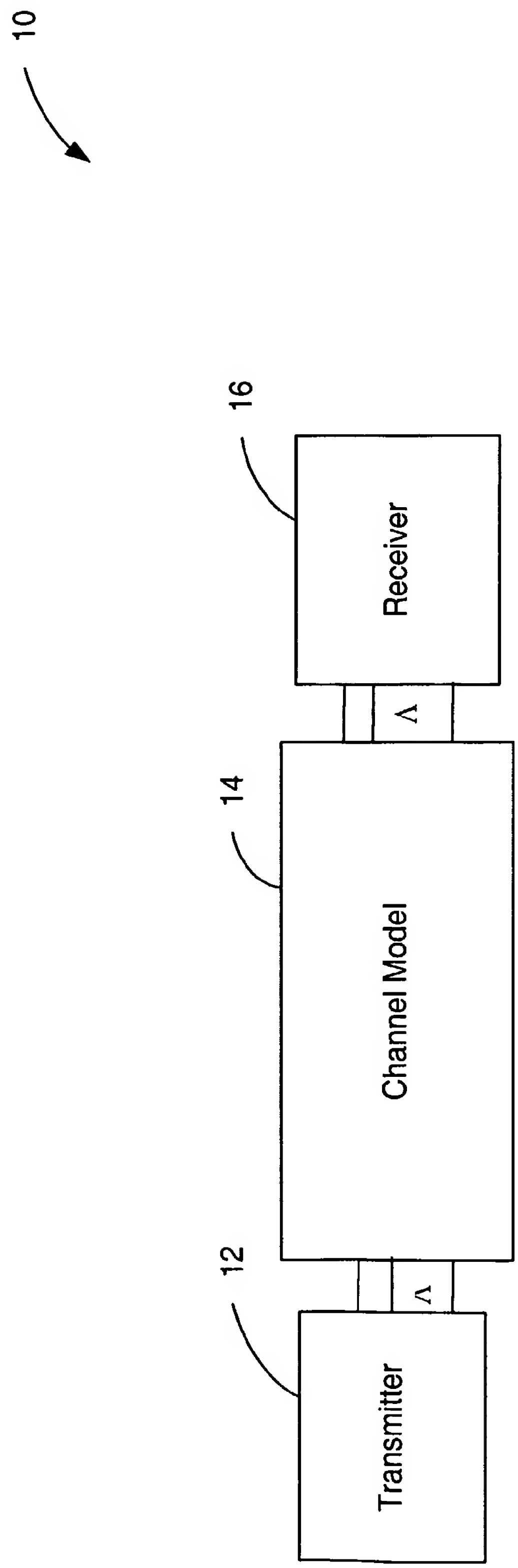


FIG. 2

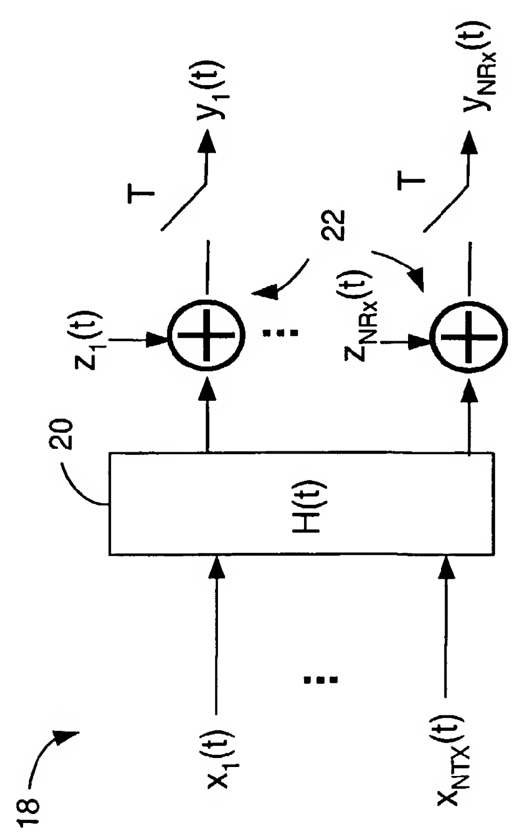

18
20
H(t)
$x_1(t)$
$x_{NTX}(t)$
$z_1(t)$
$z_{NRx}(t)$
22
T
T
$y_1(t)$
$y_{NRx}(t)$

FIG. 3

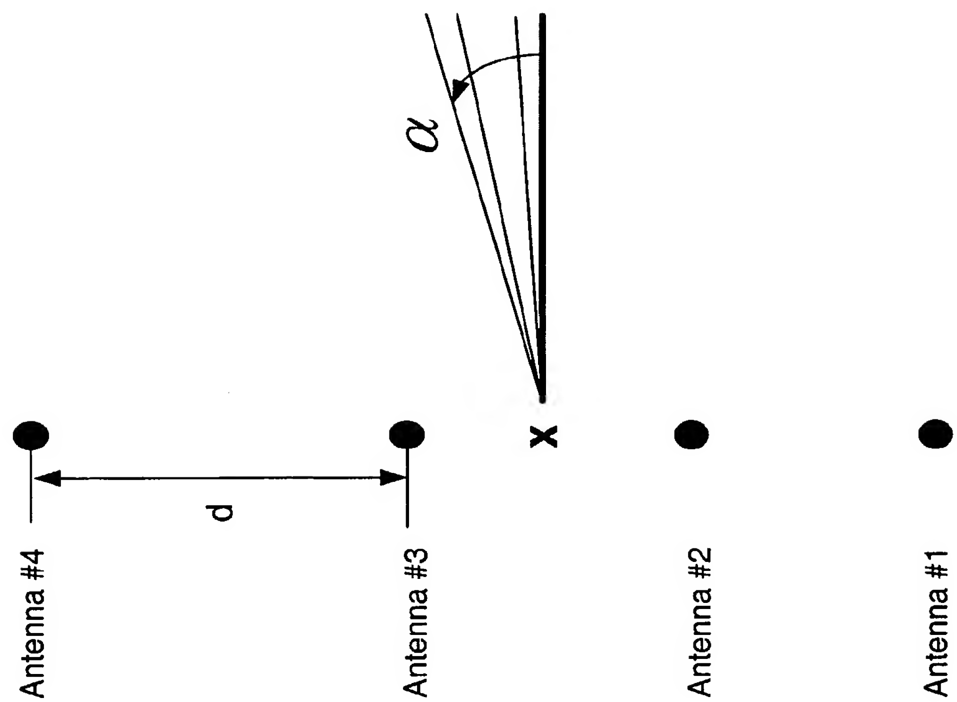

Antenna #4
Antenna #3
Antenna #2
Antenna #1
d
x
α

FIG. 4

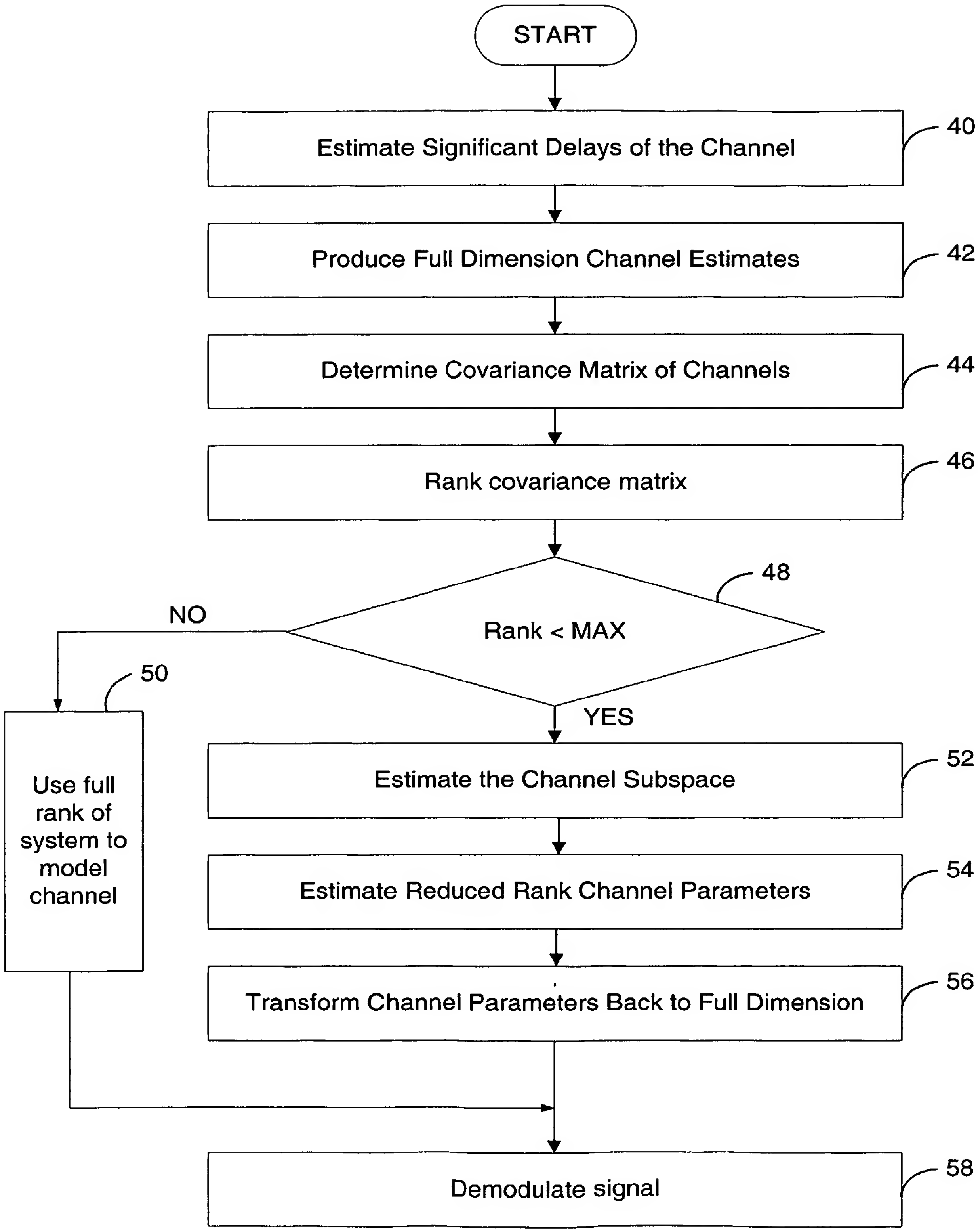

START
Estimate Significant Delays of the Channel
40
Produce Full Dimension Channel Estimates
42
Determine Covariance Matrix of Channels
44
Rank covariance matrix
46
Rank < MAX
48
NO
50
Use full rank of system to model channel
YES
Estimate the Channel Subspace
52
Estimate Reduced Rank Channel Parameters
54
Transform Channel Parameters Back to Full Dimension
56
Demodulate signal
58

FIG. 5

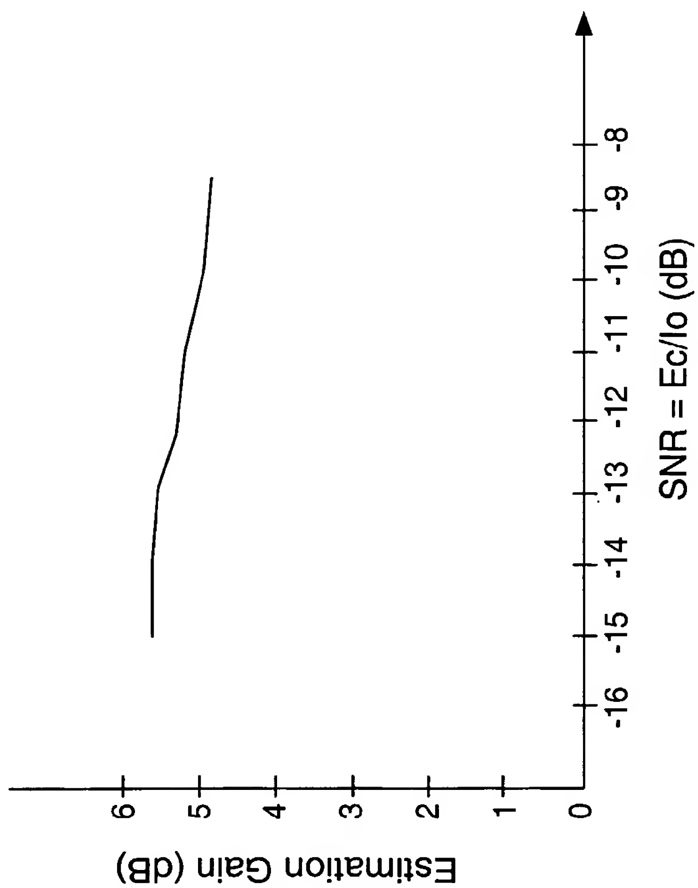


FIG. 6

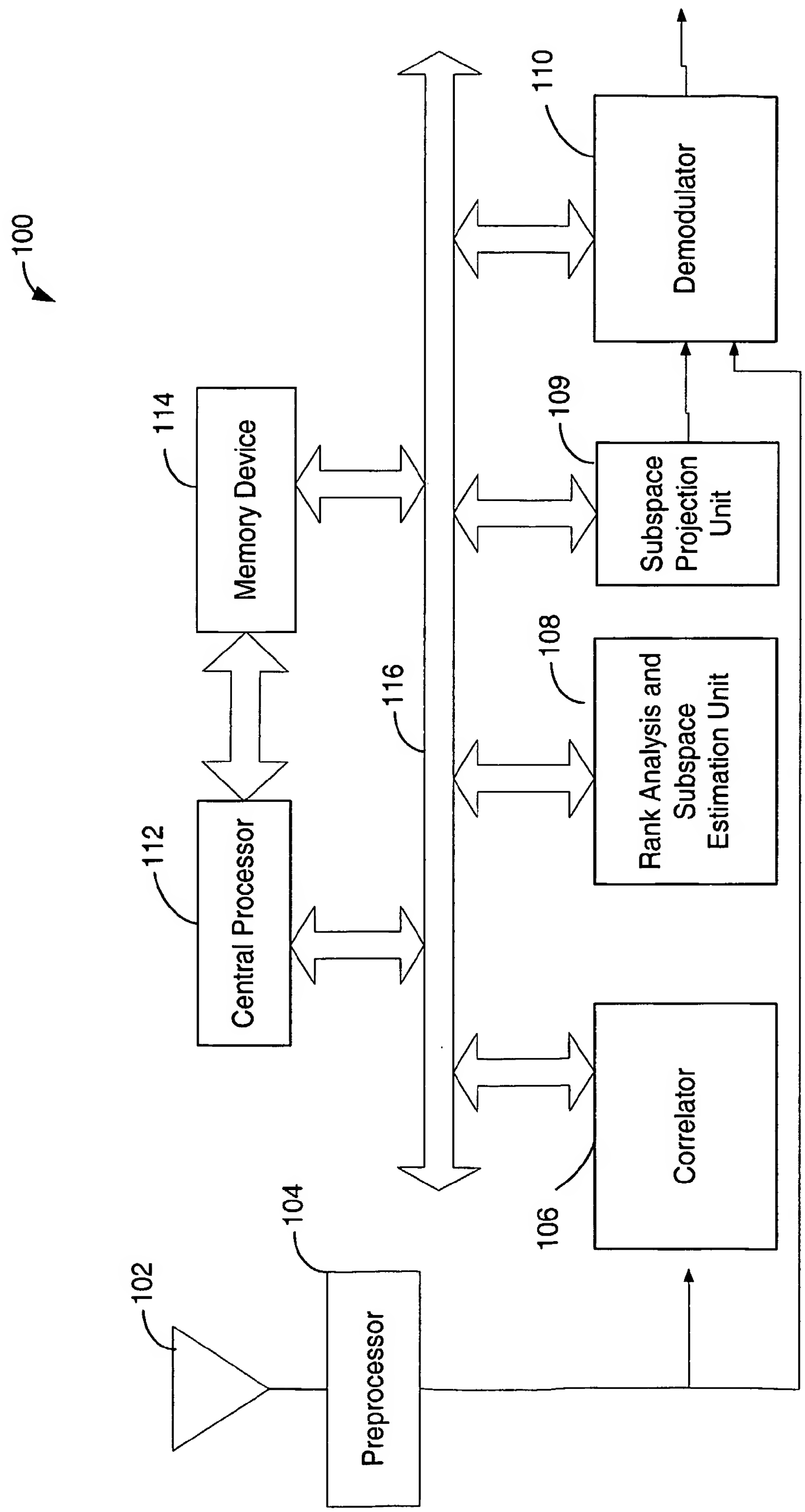
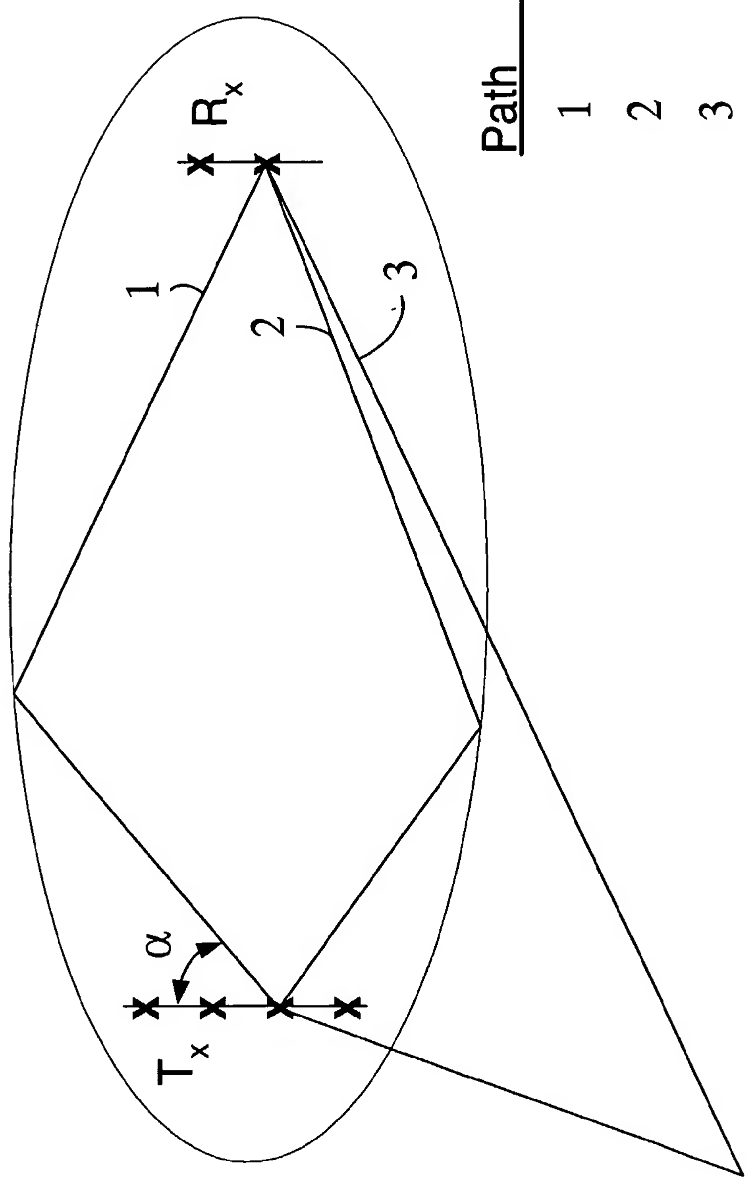

100
102
Preprocessor
104
Correlator
106
Central Processor
112
Memory Device
114
116
Rank Analysis and Subspace Estimation Unit
108
Subspace Projection Unit
109
Demodulator
110

FIG. 7



Path	Delay
1	τ_1
2	τ_1
3	τ_2

FIG. 8

INTERNATIONAL SEARCH REPORT

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B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
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Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

WPI Data, EPO-Internal, PAJ, INSPEC, COMPENDEX

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	LINDSKOG E ET AL: "REDUCED RANK CHANNEL ESTIMATION" 1999 IEEE 49TH. VEHICULAR TECHNOLOGY CONFERENCE. HOUSTON, TX, MAY 16 - 20, 1999, IEEE VEHICULAR TECHNOLOGY CONFERENCE, NEW YORK, NY: IEEE, US, vol. 2 CONF. 49, 16 May 1999 (1999-05-16), pages 1126-1130, XP000903221 ISBN: 0-7803-5566-0 section I page 1128, right-hand column --- -/--	1-27

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance
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Date of the actual completion of the international search	Date of mailing of the international search report
25 October 2002	06/11/2002

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C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	GIANCOLA D ET AL: "VARIABLE RANK RECEIVER STRUCTURES FOR LOW-RANK SPACE-TIME CHANNELS" 1999 IEEE 49TH. VEHICULAR TECHNOLOGY CONFERENCE. HOUSTON, TX, MAY 16 - 20, 1999, IEEE VEHICULAR TECHNOLOGY CONFERENCE, NEW YORK, NY: IEEE, US, vol. 1 CONF. 49, 16 May 1999 (1999-05-16), pages 65-69, XP000899198 ISBN: 0-7803-5566-0 figure 1 page 66, right-hand column page 67, left-hand column ---	1-27
A	WO 98 09381 A (UNIV LELAND STANFORD JUNIOR) 5 March 1998 (1998-03-05) abstract -----	1-27

INTERNATIONAL SEARCH REPORT

Information on patent family members

In ional Application No
PCT/US 01/50068

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO 9809381 A	05-03-1998	AU 4238697 A	19-03-1998
		CA 2302289 A1	05-03-1998
		EP 0920738 A1	09-06-1999
		EP 0931388 A2	28-07-1999
		JP 2001505723 T	24-04-2001
		WO 9809385 A2	05-03-1998
		WO 9809381 A1	05-03-1998
		WO 9809395 A1	05-03-1998
		US 6377631 B1	23-04-2002
		US 6452981 B1	17-09-2002
		US 6144711 A	07-11-2000

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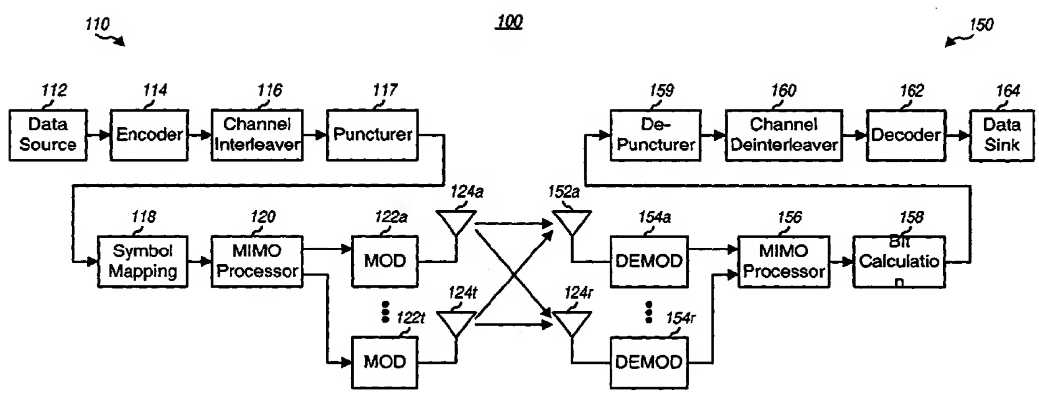
(54) Title: CODING SCHEME FOR A WIRELESS COMMUNICATION SYSTEM



(57) Abstract: Coding techniques for a (e.g., OFDM) communication system capable of transmitting data on a number of transmission channels at different information bit rates based on the channels' achieved SNR. A base code is used in combination with common or variable puncturing to achieve different coding rates required by transmission channels. The data (i.e., information bits) for a data transmission is encoded with the base code, and the coded bits for each channel (or group of channels with the similar transmission capabilities) are punctured to achieve the required coding rate. The coded bits may be interleaved (e.g., to combat fading and remove correlation between coded bits in each modulation symbol) prior to puncturing. The unpunctured codes bits are grouped into non-binary symbols (e.g., using Gray mapping). The modulation symbol may be preconditioned and prior to transmission.

WO 02/062002 A1

CODING SCHEME FOR A WIRELESS COMMUNICATION SYSTEM

BACKGROUND

I. Field

The present invention relates to data communication. More particularly, the present invention relates to a novel, flexible, and efficient coding scheme for encoding data for transmission on multiple transmission channels with different transmission capabilities.

II. Description of the Related Art

Wireless communication systems are widely deployed to provide various types of communication such as voice, data, and so on. These systems may be based on code division multiple access (CDMA), time division multiple access (TDMA), orthogonal frequency division modulation (OFDM), or some other modulation techniques. OFDM systems may provide high performance for some channel environments.

In an OFDM system, the operating frequency band is effectively partitioned into a number of "frequency subchannels", or frequency bins. Each subchannel is associated with a respective subcarrier upon which data is modulated, and may be viewed as an independent "transmission channel". Typically, the data to be transmitted (i.e., the information bits) is encoded with a particular coding scheme to generate coded bits. For a high-order modulation scheme (e.g., QPSK, QAM, and so on), the coded bits are grouped into non-binary symbols that are then used to modulate the subcarriers.

The frequency subchannels of an OFDM system may experience different link conditions (e.g., different fading and multipath effects) and may achieve different signal-to-noise-plus-interference ratio (SNR). Consequently, the number of information bits per modulation symbol (i.e., the information bit rate) that may be transmitted on each subchannel for a particular level of performance may be different from subchannel to subchannel. Moreover, the link conditions typically vary with time. As a result, the supported bit rates for the subchannels also vary with time.

The different transmission capabilities of the frequency subchannels plus the time-variant nature of the capabilities make it challenging to provide an effective coding scheme capable of encoding the supported number of

information bits/modulation symbol to provide the required coded bits for the subchannels.

Accordingly, a high performance, efficient, and flexible coding scheme that may be used to encode data for transmission on multiple subchannels with different transmission capabilities is highly desirable.

SUMMARY

Various aspects of the present invention provides efficient and effective coding techniques for a communication system capable of transmitting data on a number of "transmission channels" at different information bit rates based on the channels' achieved SNR. A number of coding/puncturing schemes may be used to generate the required coded bits (i.e., the information, tail, and parity bits, if a Turbo code is used). In a first coding/puncturing scheme, a particular base code and common puncturing is used for all transmission channels (e.g., all frequency subchannels in an OFDM system, or spatial subchannels of all frequency subchannels in an OFDM system with multiple input/multiple output antennas (MIMO), as described below). In a second coding/puncturing scheme, the same base code but variable puncturing is used for the transmission channels. The variable puncturing can be used to provide different coding rates for the transmission channels. The coding rate for each transmission channel is dependent on the information bit rate and the modulation scheme selected for the channel.

An embodiment of the invention provides a method for preparing data for transmission on a number of transmission channels in a communication system, e.g., an orthogonal frequency division modulation (OFDM) system. Each transmission channel is operable to transmit a respective sequence of modulation symbols. In accordance with the method, the number of information bits per modulation symbol supported by each transmission channel is determined (e.g., based on the channel's SNR). A modulation scheme is then identified for each transmission channel such that the determined number of information bits per modulation symbol is supported. Based on the supported number of information bits per modulation symbol and the identified modulation scheme, the coding rate for each transmission channel is determined. At least two transmission channels are associated with different coding rates because of different transmission capabilities.

Thereafter, a number of information bits is encoded in accordance with a particular encoding scheme to provide a number of coded bits. If a Turbo code

is used, a number of tail and parity bits are generated for the information bits (the coded bits include the information bits, tail bits, and parity bits). The coded bits may be interleaved in accordance with a particular interleaving scheme. For ease of implementation, the interleaving may be performed prior to puncturing. The coded bits (e.g., the tail and parity bits, if a Turbo code is used) are then punctured in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the transmission channels. The puncturing is adjusted to achieve different coding rates needed by the transmission channels. As an alternative, the puncturing may also be performed prior to interleaving.

Non-binary symbols are then formed for the transmission channels. Each non-binary symbol includes a group of interleaved and unpunctured coded bits and is mapped a respective modulation symbol. The specific number of coded bits in each non-binary symbol is dependent on the channel's modulation scheme. For a multiple-input multiple-output (MIMO) system capable of transmitting on a number of spatial subchannels for each frequency subchannel, the modulation symbols for each frequency subchannel may be pre-conditioned prior to transmission, as described below.

The invention provides methods and system elements that implement various aspects, embodiments, and features of the invention, as described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system capable of implementing various aspects and embodiments of the invention;

FIG. 2 is a diagram that graphically illustrates an OFDM transmission from one of N_T transmit antennas in the MIMO system;

FIGS. 3A and 3B are diagrams of a parallel concatenated convolutional encoder;

FIG. 3C is a diagram of an embodiment of a puncturer and multiplexer, which may be used to provide variable puncturing of coded bits;

FIGS. 4A and 4B are flow diagrams of two coding/puncturing schemes for generating the required coded bits for a data transmission, which utilize a particular base code but common and variable puncturing schemes, respectively;

FIG. 5 is a diagram of a signal constellation for 16-QAM and a specific Gray mapping scheme;

FIG. 6 is a block diagram of an embodiment of a MIMO processor;

FIG. 7 is a block diagram of an embodiment of a system capable of providing different processing for different transmissions; and

FIG. 8 is a block diagram of an embodiment of the decoding portion of a receiving system.

DETAILED DESCRIPTION OF THE SPECIFIC EMBODIMENTS

FIG. 1 is a diagram of a multiple-input multiple-output (MIMO) communication system 100 capable of implementing various aspects and embodiments of the invention. Communication system 100 can be designed to implement the coding schemes described herein. System 100 can further be operated to employ a combination of antenna, frequency, and temporal diversity to increase spectral efficiency, improve performance, and enhance flexibility. Increased spectral efficiency is characterized by the ability to transmit more bits per second per Hertz (bps/Hz) when and where possible to better utilize the available system bandwidth. Improved performance may be quantified, for example, by a lower bit-error-rate (BER) or frame-error-rate (FER) for a given link signal-to-noise-plus-interference ratio (SNR). And enhanced flexibility is characterized by the ability to accommodate multiple users having different and typically disparate requirements. These goals may be achieved, in part, by employing a high performance and efficient coding scheme, multi-carrier modulation, time division multiplexing (TDM), multiple transmit and/or receive antennas, other techniques, or a combination thereof. The features, aspects, and advantages of the invention are described in further detail below.

As shown in FIG. 1, communication system 100 includes a first system 110 in communication with a second system 150. Within system 110, a data source 112 provides data (i.e., information bits) to an encoder 114 that encodes the data in accordance with a particular coding scheme. The encoding increases the reliability of the data transmission. The coded bits are then provided to a channel interleaver 116 and interleaved (i.e., reordered) in

accordance with a particular interleaving scheme. The interleaving provides time and frequency diversity for the coded bits, permits the data to be transmitted based on an average SNR for the subchannels used for the data transmission, combats fading, and further removes correlation between coded bits used to form each modulation symbol, as described below. The interleaved bits are then punctured (i.e., deleted) to provide the required number of coded bits. The encoding, channel interleaving, and puncturing are described in further detail below. The unpunctured coded bits are then provided to a symbol mapping element 118.

In an OFDM system, the operating frequency band is effectively partitioned into a number of "frequency subchannels" (i.e., frequency bins). At each "time slot" (i.e., a particular time interval that may be dependent on the bandwidth of the frequency subchannel), a "modulation symbol" may be transmitted on each frequency subchannel. As described in further detail below, the OFDM system may be operated in a MIMO mode in which multiple (N_T) transmit antennas and multiple (N_R) receive antennas are used for a data transmission. The MIMO channel may be decomposed into N_C independent channels, with $N_C \leq N_T$ and $N_C \leq N_R$. Each of the N_C independent channels is also referred to as a "spatial subchannel" of the MIMO channel, which corresponds to a dimension. In the MIMO mode, increased dimensionality is achieved and N_C modulation symbols may be transmitted on N_C spatial subchannels of each frequency subchannel at each time slot. In an OFDM system not operated in the MIMO mode, there is only one spatial subchannel. Each frequency subchannel/spatial subchannel may also be referred to as a "transmission channel". The MIMO mode and spatial subchannel are described in further detail below.

The number of information bits that may be transmitted for each modulation symbol for a particular level of performance is dependent on the SNR of the transmission channel. For each transmission channel, symbol mapping element 118 groups a set of unpunctured coded bits to form a non-binary symbol for that transmission channel. The non-binary symbol is then mapped to a modulation symbol, which represents a point in a signal constellation corresponding to the modulation scheme selected for the transmission channel. The bit grouping and symbol mapping are performed for all transmission channels, and for each time slot used for data transmission. The modulation symbols for all transmission channels are then provided to a MIMO processor 120.

Depending on the particular "spatial" diversity being implemented (if any), MIMO processor 120 may demultiplex, pre-condition, and combine the received modulation symbols. The MIMO processing is described in further detail below. For each transmit antenna, MIMO processor 120 provides a stream of modulation symbol vectors, one vector for each time slot. Each modulation symbol vector includes the modulation symbols for all frequency subchannels for a given time slot. Each stream of modulation symbol vectors is received and modulated by a respective modulator (MOD) 122, and transmitted via an associated antenna 124.

In the embodiment shown in FIG. 1, receiving system 150 includes a number of receive antennas 152 that receive the transmitted signals and provide the received signals to respective demodulators (DEMOD) 154. Each demodulator 154 performs processing complementary to that performed at modulator 122. The demodulated symbols from all demodulators 154 are provided to a MIMO processor 156 and processed in a complementary manner as that performed at MIMO processor 120. The received symbols for the transmission channels are then provided to a bit calculation unit 158 that performs processing complementary to that performed by symbol mapping element 118 and provides values indicative of the received bits. Erasures (e.g., zero value indicatives) are then inserted by a de-puncturer 159 for coded bits punctured at system 110. The de-punctured values are then deinterleaved by a channel deinterleaver 160 and further decoded by a decoder 162 to generate decoded bits, which are then provided to a data sink 164. The channel deinterleaving, de-puncturing, and decoding are complementary to the channel interleaving, puncturing, and encoding performed at the transmitter.

FIG. 2 is a diagram that graphically illustrates an OFDM transmission from one of N_T transmit antennas in a MIMO system. In FIG. 2, the horizontal axis represents time and the vertical axis represents frequency. In this specific example, the transmission channel includes 16 frequency subchannels and is used to transmit a sequence of OFDM symbols, with each OFDM symbol covering all 16 frequency subchannels. A time division multiplexing (TDM) structure is also illustrated in which the data transmission is partitioned into time slots, with each time slot having a particular duration. For the example shown in FIG. 2, the time slot is equal to the length of one modulation symbol.

The available frequency subchannels may be used to transmit signaling, voice, packet data, and so on. In the specific example shown in FIG. 2, the modulation symbol at time slot 1 corresponds to pilot data, which may be periodically transmitted to assist the receiver units synchronize and perform

channel estimation. Other techniques for distributing pilot data over time and frequency may also be used. Transmission of the pilot modulation symbol typically occurs at a particular rate, which is usually selected to be fast enough to permit accurate tracking of variations in the communication link.

The time slots not used for pilot transmissions can be used to transmit various types of data. For example, frequency subchannels 1 and 2 may be reserved for the transmission of control and broadcast data to the receiver units. The data on these subchannels is generally intended to be received by all receiver units. However, some of the messages on the control channel may be user specific, and may be encoded accordingly.

Voice data and packet data may be transmitted in the remaining frequency subchannels. For the example shown, subchannel 3 at time slots 2 through 9 is used for voice call 1, subchannel 4 at time slots 2 through 9 is used for voice call 2, subchannel 5 at time slots 5 through 9 is used for voice call 3, and subchannel 6 at time slots 7 through 9 is used for voice call 5.

The remaining available frequency subchannels and time slots may be used for transmissions of traffic data. A particular data transmission may occur over multiple subchannels and/or multiple time slots, and multiple data transmissions may occur within any particular time slot. A data transmission may also occur over non-contiguous time slots.

In the example shown in FIG. 2, data 1 transmission uses frequency subchannels 5 through 16 at time slot 2 and subchannels 7 through 16 at time slot 7, data 2 transmission uses subchannels 5 through 16 at time slots 3 and 4 and subchannels 6 through 16 at time slots 5, data 3 transmission uses subchannels 6 through 16 at time slot 6, data 4 transmission uses subchannels 7 through 16 at time slot 8, data 5 transmission uses subchannels 7 through 11 at time slot 9, and data 6 transmission uses subchannels 12 through 16 at time slot 9. Data 1 through 6 transmissions can represent transmissions of traffic data to one or more receiver units.

To provide the transmission flexibility and achieve high performance and efficiency, each frequency subchannel at each time slot for each transmit antenna may be viewed as an independent unit of transmission (a modulation symbol) that may be used to transmit any type of data such as pilot, signaling, broadcast, voice, traffic data, some other data type, or a combination thereof. Flexibility, performance, and efficiency may further be achieved by allowing for independence among the modulation symbols, as described below. For example, each modulation symbol may be generated from a modulation

scheme (e.g., M-PSK, M-QAM, or some other scheme) that results in the best use of the resource at that particular time, frequency, and space.

MIMO System

In a terrestrial communications system (e.g., a cellular system, a broadcast system, a multi-channel multi-point distribution system (MMDS) system, and others), an RF modulated signal from a transmitter unit may reach the receiver unit via a number of transmission paths. The characteristics of the transmission paths typically vary over time due to a number of factors. If more than one transmit or receive antenna is used, and if the transmission paths between the transmit and receive antennas are linearly independent (i.e., one transmission is not formed as a linear combination of the other transmissions), which is generally true to at least an extent, then the likelihood of correctly receiving the transmitted signal increases as the number of antennas increases. Generally, as the number of transmit and receive antennas increases, diversity increases and performance improves.

A MIMO communication system such as the one shown in FIG. 1 employs antennas at both the transmit and receive ends of the communication link. These transmit and receive antennas may be used to provide various forms of "spatial diversity", including "transmit" diversity and "receive" diversity. Spatial diversity is characterized by the use of multiple transmit antennas and one or more receive antennas. Transmit diversity is characterized by the transmission of data over multiple transmit antennas. Typically, additional processing is performed on the data transmitted from the transmit antennas to achieved the desired diversity. For example, the data transmitted from different transmit antennas may be delayed or reordered in time, coded and interleaved across the available transmit antennas, and so on. Receive diversity is characterized by the reception of the transmitted signals on multiple receive antennas, and diversity is achieved by simply receiving the signals via different signal paths.

Spatial diversity may be used to improve the reliability of the communication link with or without increasing the link capacity. This may be achieved by transmitting or receiving data over multiple paths via multiple antennas. Spatial diversity may be dynamically selected based on the characteristics of the communication link to provide the required performance. For example, higher degree of spatial diversity may be provided for some types of communication (e.g., signaling), for some types of services (e.g., voice), for

some communication link characteristics (e.g., low SNR), or for some other conditions or considerations.

The data may be transmitted from multiple antennas and/or on multiple frequency subchannels to obtain the desired diversity. For example, data may be transmitted on: (1) one subchannel from one antenna, (2) one subchannel (e.g., subchannel 1) from multiple antennas, (3) one subchannel from all N_T antennas, (4) a set of subchannels (e.g., subchannels 1 and 2) from one antenna, (5), a set of subchannels from multiple antennas, (6) a set of subchannels from all N_T antennas, or (7) a set of channels from a set of antennas (e.g., subchannel 1 from antennas 1 and 2 at one time slot, subchannels 1 and 2 from antenna 2 at another time slot, and so on). Thus, any combination of subchannels and antennas may be used to provide antenna and frequency diversity.

In the MIMO communication system, the multi-input multi-output channel can be decomposed into a set of N_C independent spatial subchannels. The number of such spatial subchannels is less than or equal to the lesser of the number of the transmitting antennas and the number of receiving antennas (i.e., $N_C \leq N_T$ and $N_C \leq N_R$). If **H** is the N_R x N_T matrix that gives the channel response for the N_T transmit antennas and the N_R receive antennas at a specific time, and $\underline{\mathbf{x}}$ is the N_T-vector inputs to the channel, then the received signal can be expressed as:

$$\underline{\mathbf{y}} = \mathbf{H}\underline{\mathbf{x}} + \underline{\mathbf{n}} \ ,$$

where $\underline{\mathbf{n}}$ is an N_R-vector representing noise plus interference. In one embodiment, the eigenvector decomposition of the Hermitian matrix formed by the product of the channel matrix with its conjugate-transpose can be expressed as:

$$\mathbf{H}^*\mathbf{H} = E\Lambda E^* \ ,$$

where the symbol "*" denotes conjugate-transpose, E is the eigenvector matrix, and Λ is a diagonal matrix of eigenvalues, both of dimension $N_T x N_T$.

The transmitter converts (i.e., pre-conditions) a set of N_T modulation symbols $\underline{\mathbf{b}}$ using the eigenvector matrix E. The transmitted modulation symbols from the N_T transmit antennas can be expressed as:

$$\underline{\mathbf{x}} = E\underline{\mathbf{b}} \ .$$

For all antennas, the pre-conditioning of the modulation symbols can be achieved by a matrix multiply operation expressed as:

$$\begin{bmatrix} x_1 \\ x_2 \\ \mathrm{M} \\ x_{N_T} \end{bmatrix} = \begin{bmatrix} e_{11}, & e_{12}, & & e_{1N_T} \\ e_{21}, & e_{22}, & & e_{2N_T} \\ & & & \\ e_{N_T1}, & e_{N_T1}, & & e_{N_TN_T} \end{bmatrix} \cdot \begin{bmatrix} b_1 \\ b_2 \\ \mathrm{M} \\ b_{N_T} \end{bmatrix} \qquad \text{Eq (1)}$$

where b_1, b_2, ... and b_{NT} are respectively the modulation symbols for a particular frequency subchannel at transmit antennas 1, 2, ... N_T, where each modulation symbol can be generated using, for example, M-PSK, M-QAM, and so on, as described below;

E = is the eigenvector matrix related to the transmission characteristics from transmit antennas to the receive antennas; and

x_1, x_2, ... x_{NT} are the pre-conditioned modulation symbols, which can be expressed as:

$$x_1 = b_1 \cdot e_{11} + b_2 \cdot e_{12} + \ \ldots \ + b_{N_T} \cdot e_{1N_T} \ ,$$

$$x_2 = b_1 \cdot e_{21} + b_2 \cdot e_{22} + \ \ldots \ + b_{N_T} \cdot e_{2N_T} \ , \text{ and}$$

$$x_{N_T} = b_1 \cdot e_{N_T1} + b_2 \cdot e_{N_T2} + \ \ldots \ + b_{N_T} \cdot e_{N_TN_T} \ .$$

The received signal may be expressed as:

$$\underline{\mathbf{y}} = \mathbf{HE}\underline{\mathbf{b}} + \underline{\mathbf{n}} \ .$$

The receiver performs a channel-matched-filter operation, followed by multiplication by the right eigenvectors. The result of the channel-matched-filter operation is the vector $\underline{\mathbf{z}}$, which can be expressed as:

$$\underline{\mathbf{z}} = E^*\mathbf{H}^*\mathbf{H}E\underline{\mathbf{b}} + E^*\mathbf{H}^*\underline{\mathbf{n}} = \mathbf{\Lambda}\underline{\mathbf{b}} + \underline{\hat{\mathbf{n}}} \ ,$$

where the new noise term has covariance that can be expressed as:

$$E(\underline{\hat{n}}\underline{\hat{n}}^*) = E(E^*\mathbf{H}^*\underline{\mathbf{nn}}^*\mathbf{H}E) = E^*\mathbf{H}^*\mathbf{H}E = \Lambda \ ,$$

i.e., the noise components are independent and have variance given by the eigenvalues. The SNR of the i[th] component of $\underline{\mathbf{z}}$ is λ_i, the i[th] diagonal element of $\mathbf{\Lambda}$.

An embodiment of the MIMO processing is described in further detail below and in U.S Patent Application Serial No. 09/532,491, entitled "HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM EMPLOYING MULTI-CARRIER MODULATION," filed March 22, 2000,

assigned to the assignee of the present application and incorporated herein by reference.

Each of the N_C spatial subchannels in the MIMO channel as described in the above embodiment is also referred to as an eigenmode if these channels are independent of each other. For the MIMO mode, one modulation symbol can be transmitted on each of the eigenmodes in each frequency subchannel. Since the SNR may be different for each eigenmode, the number of bits that may be transmitted over each eigenmode may also be different. As noted above, each eigenmode of each frequency subchannel is also referred to as a transmission channel.

In other embodiments, the spatial subchannels can be created differently. For example, a spatial subchannel can be defined as the transmissions from one transmitter antenna to all of the receiver antennas.

As used herein, the MIMO mode includes full channel state information (full-CSI) and partial-CSI processing modes. For both full-CSI and partial-CSI, additional transmission paths are provided via spatially separable subchannels. Full-CSI processing utilizes eigenmodes, as described above. Partial-CSI processing does not utilize eigenmodes, and may involve providing to the transmitter unit (e.g., via feeding back on the reverse link) the SNR for each transmission channel (i.e., receive diversity port), and coding accordingly based on the received SNR.

A number of formulations may be utilized at the receiver unit to provide the requisite information for partial-CSI, including linear and non-linear forms of zero-forcing, channel correlation matrix inversion (CCMI), and minimum mean square error (MMSE), as is known in the art. For example, the derivation of SNRs for a non-linear zero-forcing (partial-CSI) MIMO case is described by P.W. Wolniansky *et al.* in a paper entitled "V-BLAST: An Architecture for Realizing Very High Data Rates Over the Rich-Scattering Wireless Channel," Proc. IEEE ISSSE-98, Pisa, Italy, Sept. 30, 1998, and incorporated herein by reference. The eigenvalues from a MIMO formulation are related to the SNRs of the eigenmodes for the full-CSI case. Non-MIMO cases can use an assortment of methods, as is known in the art.

Each transmission channel is associated with a SNR that may be known to both the transmitter and receiver. In this case, the modulation and coding parameters of each modulation symbol can be determined based on the SNR of the corresponding transmission channel. This allows for efficient use of the available frequency subchannels and eigenmodes.

Table 1 lists the number of information bits that may be transmitted in each modulation symbol for a particular level of performance (e.g., 1% frame-error rate, or % FER) for various SNR ranges. For each SNR range, Table 1 also lists a particular modulation scheme selected for use with that SNR range, the number of coded bits that may be transmitted for each modulation symbol for the selected modulation scheme, and the coding rate used to obtain the required number of coded bits/modulation symbol given the supported number of information bits/modulation symbol.

Table 1 lists one combination of modulation scheme and coding rate for each SNR range. The supported bit rate for each transmission channel may be achieved using any one of a number of possible combinations of coding rate and modulation scheme. For example, one information bit per symbol may be achieved using (1) a coding rate of 1/2 and QPSK modulation, (2) a coding rate of 1/3 and 8-PSK modulation, (3) a coding rate of 1/4 and 16-QAM, or (4) some other combination of coding rate and modulation scheme. In Table 1, QPSK, 16-QAM, and 64-QAM are used for the listed SNR ranges. Other modulation schemes such as 8-PSK, 32-QAM, 128-QAM, and so on, may also be employed and are within the scope of the invention.

Table 1

SNR Range	# of Information Bits/Symbol	Modulation Symbol	# of Coded Bits/Symbol	Coding Rate
1.5 – 4.4	1	QPSK	2	1/2
4.4 – 6.4	1.5	QPSK	2	3/4
6.4 – 8.35	2	16-QAM	4	1/2
8.35 – 10.4	2.5	16-QAM	4	5/8
10.4 – 12.3	3	16-QAM	4	3/4
12.3 – 14.15	3.5	64-QAM	6	7/12
14.15 – 15.55	4	64-QAM	6	2/3
15.55 – 17.35	4.5	64-QAM	6	3/4
> 17.35	5	64-QAM	6	5/6

For clarity, various aspects of the invention are described for an OFDM system and, in many instances, for an OFDM system operating in a MIMO mode. However, the encoding and processing techniques described herein may generally be applied to various communication systems such as, for example, (1) an OFDM system operating without MIMO, (2) a MIMO system operating without OFDM (i.e., operating based on a single frequency

subchannel, i.e., a single RF carrier, but multiple spatial subchannels), (3) a MIMO system operating with OFDM, and (4) others. OFDM is simply one technique for subdividing a wideband channel into a number of orthogonal frequency subchannels.

Encoding

FIG. 3A is a block diagram of an embodiment of a parallel concatenated convolutional encoder 114x, which is often referred to as a Turbo encoder. Turbo encoder 114x represents one implementation of the forward error correction (FEC) portion of encoder 114 in FIG. 1 and may be used to encode data for transmission over one or more transmission channels.

The encoding within encoder 114 may include error correction coding or error detection coding, or both, which are used to increase the reliability of the link. The encoding may include, for example, cyclic redundancy check (CRC) coding, convolutional coding, Turbo coding, Trellis coding, block coding (e.g., Reed-Solomon coding), other types of coding, or a combination thereof. For a wireless communication system, a packet of data may be initially encoded with a particular CRC code, and the CRC bits are appended to the data packet. Additional overhead bits may also be appended to the data packet to form a formatted data packet, which is then encoded with a convolutional or Turbo code. As used herein, "information bits" refer to bits provided to the convolutional or Turbo encoder, including transmitted data bits and bits used to provide error detection or correction capability for the transmitted bits.

As shown in FIG. 3A, Turbo encoder 114x includes two constituent encoders 312a and 312b, and a code interleaver 314. Constituent encoder 312a receives and encodes the information bits, x, in accordance with a first constituent code to generate a first sequence of tail and parity bits, y. Code interleaver 314 receives and interleaves the information bits in accordance with a particular interleaving scheme. Constituent encoder 312b receives and encodes the interleaved bits in accordance with a second constituent code to generate a second sequence of tail and parity bits, z. The information bits, tail bits, and parity bits from encoders 312a and 312b are provided to the next processing element (channel interleaver 116).

FIG. 3B is a diagram of an embodiment of a Turbo encoder 114y, which is one implementation of Turbo encoder 114x and may also be used within encoder 114 in FIG. 1. In this example, Turbo encoder 114y is a rate 1/3 encoder that provides two parity bits, y and z, for each information bit x.

In the embodiment shown in FIG. 3B, each constituent encoder 322 of Turbo encoder 114y implements the following transfer function for the constituent code:

$$G(D) = \begin{bmatrix} 1 & \dfrac{n(D)}{d(D)} \end{bmatrix} ,$$

where

$$n(D) = 1 + D + D^3 \quad , \textit{ and}$$

$$d(D) = 1 + D^2 + D^3$$

Other constituent codes may also be used and are within the scope of the invention.

Each constituent encoder 322 includes a number of series coupled delay elements 332, a number of modulo-2 adders 334, and a switch 336. Initially, the states of delay elements 332 are set to zeros and switch 336 is in the up position. Then, for each information bit in a data packet, adder 334a performs modulo-2 addition of the information bit with the output bit from adder 334c and provides the result to delay element 332a. Adder 334b receives and performs modulo-2 addition of the bits from adder 334a and delay elements 332a and 332c, and provides the parity bit y. Adder 334c performs modulo-2 addition of the bits from delay elements 332b and 332c.

After all N information bits in the data packet have been encoded, switch 336 is moved to the down position and three zero ("0") bits are provided to the constituent encoder 322a. Constituent encoder 322a then encodes the three zero bits and provides three tail systematic bits and three tail parity bits.

For each packet of N information bits, constituent encoder 322a provides N information bits x, the first three tail systematic bits, N parity bits y, and the first three tail parity bits, and constituent encoder 322b provides the second three tail systematic bits, N parity bits z, and the last three tail parity bits. For each packet, encoder 114y provides N information bits, six tail systematic bits, N+3 parity bits from encoder 322a, and N+3 parity bits from encoder 322b.

Code interleaver 314 may implement any one of a number of interleaving schemes. In one specific interleaving scheme, the N information bits in the packet are written, by row, into a 2^5-row by 2^n-column array, where n is the smallest integer such that $N \leq 2^{5+n}$. The rows are then shuffled according to a bit-reversal rule. For example, row 1 ("00001") is swapped with row 16 ("10000"), row 3 ("00011") is swapped with row 24 ("11000"), and so on. The bits

within each row are then permutated (i.e., rearranged) according to a row-specific linear congruential sequence (LCS). The LCS for row k may be defined as $x_k(i+1) = \{x_k(i) + c_k\} \bmod 2^n$, where $i = 0, 1, \ldots 2^n-1$, $x_k(0) = c_k$, and c_k is a specific value selected for each row and is further dependent on the value for n. For permutation in each row, the i^{th} bit in the row is placed in location x(i). The bits in code interleaver 314 are then read out by column.

The above LCS code interleaving scheme is described in further detail in commonly assigned U.S. Patent Application Serial No. 09/205,511, entitled "TURBO CODE INTERLEAVER USING LINEAR CONGRUENTIAL SEQUENCES," filed December 4, 1998, and in a document entitled "C.S0002-A-1 Physical Layer Standard for cdma2000 Spread Spectrum Systems" (hereinafter referred to as the cdma2000 standard), both of which are incorporated herein by reference.

Other code interleaver may also be used and are within the scope of the invention. For example, a random interleaver or a symmetrical-random (S-random) interleaver may also be used instead of the linear congruential sequence interleaver described above.

For clarity, the data coding is specifically described based on a Turbo code. Other coding schemes may also be used and are within the scope of the invention. For example, the data may be coded with a convolutional code, a block code, a concatenated code comprised of a combination of block, convolutional, and/or Turbo codes, or some other code. The data may be coded in accordance with a "base" code, and the coded bits may thereafter be processed (e.g., punctured) based on the capabilities of the transmission channels used to transmit the data.

Channel Interleaving

Referring back to FIG. 1, the coded bits from encoder 114 are interleaved by channel interleaver 116 to provide temporal and frequency diversity against deleterious path effects (e.g., fading). Moreover, since coded bits are subsequently grouped together to form non-binary symbols that are then mapped to modulation symbols, the interleaving further ensures that the coded bits that form each modulation symbol are not located close to each other (temporally). For static additive white Gaussian noise (AWGN) channels, the channel interleaving is less critical when a Turbo encoder is also employed, since the code interleaver effectively performs similar functions.

Various interleaving schemes may be used for the channel interleaver. In one interleaving scheme, the coded bits (i.e., the information, tail, and parity

bits) for each packet are written (linearly) to rows of memory. The bits in each row may then be permutated (i.e., rearranged) based on (1) a bit-reversal rule, (2) a linear congruential sequence (such as the one described above for the code interleaver), (3) a randomly generated pattern, (4) or a permutation pattern generated in some other manner. The rows are also permutated in accordance with a particular row permutation pattern. The permutated coded bits are then retrieved from each column and provided to puncturer 117.

In an embodiment, the channel interleaving is performed individually for each bit stream in a packet. For each packet, the information bits x, the tail and parity bits y from the first constituent encoder, and the tail and parity bits z from the second constituent encoder may be interleaved by three separate interleavers, which may employ the same or different channel interleaving schemes. This separate interleaving allows for flexible puncturing on the individual bit streams.

The interleaving interval may be selected to provide the desired temporal and frequency diversity. For example, coded bits for a particular time period (e.g., 10 msec, 20 msec, or some other) and/or for a particular number of transmission channels may be interleaved.

Puncturing

As noted above, for an OFDM communication system, the number of information bits that may be transmitted for each modulation symbol is dependent on the SNR of the transmission channel used to transmit the modulation symbol. And for an OFDM system operated in the MIMO mode, the number of information bits that may be transmitted for each modulation symbol is dependent on the SNR of the frequency subchannel and spatial subchannel used to transmit the modulation symbol.

In accordance with an aspect of the invention, a number of coding/puncturing schemes may be used to generate the coded bits (i.e., information, tail, and parity bits) for transmission. In a first coding/puncturing scheme, a particular base code and common puncturing is applied for all transmission channels. In a second coding/puncturing scheme, the same base code but variable puncturing is applied for the transmission channels. The variable puncturing is dependent on the SNR of the transmission channels.

FIG. 4A is a flow diagram of an embodiment for generating the required coded bits for a data transmission, which employs the base code and common puncturing scheme. Initially, the SNR for each transmission channel (i.e., each eigenmode of each frequency subchannel) is determined, at step 412. For an

OFDM system not operated in the MIMO mode, only one eigenmode is supported and thus only one SNR is determined for each frequency subchannel. The SNR for each transmission channel may be determined based on the transmitted pilot reference or via some other mechanism.

At step 414, the number of information bits per modulation symbol supported by each transmission channel is determined based on its SNR. A table that associates a range of SNR with each specific number of information bits/modulation symbol, such as Table 1, may be used. However, finer quantization than the 0.5-bit step size for the information bits shown in Table 1 may be used. A modulation scheme is then selected for each transmission channel such that the number of information bits/modulation symbol can be transmitted, at step 416. The modulation scheme may also be selected to take into account other factors (e.g., coding complexity), as described in further detail below.

At step 418, the total number of information bits that may be transmitted in each time slot for all transmission channels is determined. This can be achieved by summing the number of information bits/modulation symbol determined for all transmission channels. Similarly, the total number of coded bits that may be transmitted in each time slot for all transmission channels is determined, at step 420. This can be achieved by determining the number of coded bits/modulation symbol for each modulation scheme selected in step 416, and summing the number of coded bits for all transmission channels.

At step 422, the total number of information bits determined in step 418 is encoded with a particular encoder. If a Turbo encoder is used, the tail bits and parity bits generated by the encoder are punctured to obtain the total number of coded bits determined in step 420. The unpunctured coded bits are then grouped into non-binary symbols, which are then mapped to modulation symbols for the transmission channels, at step 426.

The first coding/puncturing scheme is relatively simple to implement since the same base code and puncturing scheme are used for all transmission channels. The modulation symbol for each transmission channel represents a point in a signal constellation corresponding to the modulation scheme selected for that transmission channel. If the distribution of the SNR for the transmission channels is widespread, the distance between the constellation points relative to the noise variance for different signal constellations will vary widely. This may then impact the performance of the system.

FIG. 4B is a flow diagram of an embodiment for generating the required coded bits for a data transmission, which employs the same base code but

variable puncturing scheme. Initially, the SNR for each transmission channel is determined, at step 432. In an embodiment, transmission channels with insufficient SNR are omitted from use for data transmission (i.e., no data is transmitted on poor transmission channels). The number of information bits per modulation symbol supported by each transmission channel is then determined based on its SNR, at step 434. A modulation scheme is next selected for each transmission channel such that the number of information bits/modulation symbol can be transmitted, at step 436. Steps 432, 434, and 436 in FIG. 4B correspond to steps 412, 414, and 416 in FIG. 4A.

At step 438, the transmission channels belonging to the same SNR range are grouped into a segment. Alternatively, ranges can be defined for the number of information bits per modulation symbol (e.g., range 1 covering 1.0 to 1.5 information bits/modulation symbol, range 2 covering 1.5 to 2.0 information bits/modulation symbol, and so on). In this case, transmission channels having number of information bits per modulation symbol within the same range are grouped into a segment.

Each segment includes K_i transmission channels, where K_i can be any integer one or greater. The total number of information bits and total number of coded bits that can be transmitted in each segment are then determined, at step 440. For example, segment i may include K_i transmission channels, each of which may support transmission of N_i information bits/modulation symbol and P_i tail and parity bits/modulation symbol. For each time slot, the total number of information bits that may be transmitted in segment i can be computed as $K_i \cdot N_i$, the total number of tail and parity bits that may be transmitted can be computed as $K_i \cdot P_i$, and the total number of coded bits may be computed as $K_i(N_i + P_i)$.

At step 442, the information bits to be transmitted in each time slot for all segments, which may be computed as $\sum_i K_i N_i$, are encoded with a particular encoder (e.g., a rate 1/3 Turbo encoder such at the one shown in FIG. 3B). At step 444, N_i information bits and N_i/R parity and tail bits are assigned to each transmission channel of segment i, where R is the coding rate of the encoder. The N_i/R parity and tail bits are then punctured to obtain the P_i parity and tail bits required for each transmission channel of the segment, at step 446. At step 448, the N_i information bits and the P_i parity and tail bits for each transmission channel of segment i are mapped to a modulation symbol for the transmission channel.

The second coding/puncturing scheme may provide improved performance over the first scheme, especially if the distribution of SNR for the transmission channels is widespread. Since different modulation schemes and coding rate may be used for different transmission channels, the number of bits transmitted on each transmission channel is typically communicated from the receiver to the transmitter on the reverse link.

Table 1 shows the quantization of the number of information bits/modulation symbol using 0.5-bit step size. The quantization granularity may be reduced (i.e., to be finer than 0.5-bit) if each segment (and not each transmission channel) is required to support an integer number of information bits. If $K_i \cdot N_i$ is required to be an integer, a larger integer value for K_i allows for a smaller step size for N_i. The quantization granularity may be further reduced if the quantization is allowed to be carried from segment to segment. For example, if one bit needs to be rounded-off in one segment, one bit may be rounded-up in the next segment, if appropriate. The quantization granularity may also be reduced if the quantization is allowed to be carried over multiple time slots.

To support an OFDM system (especially one operated in the MIMO mode) whereby different SNR may be achieved for the transmission channels, a flexible puncturing scheme may be used in conjunction with a common base encoder (e.g., a rate 1/3 Turbo encoder) to achieve the necessary coding rates. This flexible puncturing scheme may be used to provide the necessary number of tail and parity bits for each segment. For a high coding rate in which more tail and parity bits are punctured than retained, the puncturing may be efficiently achieved by retaining the required number of tail and parity bits as they are generated by the encoder and discarding the others.

As an example, a segment may include 20 16-QAM modulation symbols and has a SNR that supports transmission of 2.75 information bits/modulation symbol. For this segment, 55 information bits (55 = 20x2.75) may be transmitted in 20 modulation symbols. Each 16-QAM modulation symbol is formed with four coded bits, and 80 coded bits are needed for 20 modulation symbols. The 55 information bits may be encoded with a rate 1/3 encoder to generate 122 tail and parity bits and 55 information bits. These 122 tail and parity bits may be punctured to provide the 35 tail and parity bits required for the segment, which in combination with the 55 information bits comprise the 80 coded bits.

Referring back to FIG. 1, puncturer 117 receives the interleaved information and parity bits from channel interleaver 116, punctures (i.e.,

deletes) some of the tail and parity bits to achieve the desired coding rate(s), and multiplexes the unpunctured information, tail, and parity bits into a sequence of coded bits. The information bits (which are also referred to as systematic bits) may also be punctured along with the tail and parity bits, and this is within the scope of the invention.

FIG. 3C is a diagram of an embodiment of a puncturer 117x, which may be used to provide variable puncturing of coded bits. Puncturer 117x is one implementation of puncturer 117 in FIG. 1. Using a set of counters, puncturer 117x performs puncturing to retain P_i tail and parity bits out of Q_i tail and parity bits generated by the encoder for segment i.

Within puncturer 117x, the interleaved tail and parity bits y_{INT} and z_{INT} from the two constituent encoders of the Turbo encoder are provided to two inputs of a switch 342. Switch 342 provides either the y_{INT} tail and parity bits or the z_{INT} tail and parity bits to line 343, depending on a control signal from a toggle unit 348. Switch 342 ensures that the tail and parity bits from the two constituent encoders are evenly selected by alternating between the two tail and parity bit streams.

A first counter 352 performs modulo-Q addition and wraps around after its content reaches beyond Q-1. A second counter 354 counts (by one) the Q tail and parity bits. For each segment, both counters 352 and 354 are initially set to zero, switch 342 is in the up position, and the first tail or parity bit y_{INT0} is provided from multiplexer 346 by closing a switch 344 and appropriately controlling the multiplexer. For each subsequent clock cycle, counter 352 is incremented by P and counter 354 is incremented by one. The value of counter 352 is provided to a decision unit 356. If counter 352 experiences a modulo-Q operation (i.e., the content of counter 352 wraps around), the tail or parity bit on line 343 is provided through switch 344 to multiplexer 346, which then provides the tail or parity bit as an output coded bit. Each time a tail or parity bit is provided from multiplexer 346, toggle unit 348 toggles the state of the control signal, and the other tail and parity bit stream is provided to line 343. The process continues until all Q_i tail and parity bits in the segment are exhausted, as indicated by comparison unit 358.

Other puncturing patterns may also be used and are within the scope of the invention. To provide good performance, the number of tail and parity bits to be punctured should be balanced between the two constituent codes (i.e., approximately equal number of y_{INT} and z_{INT} tail and parity bits are selected) and the unpunctured bits should be distributed relatively evenly over the code block for each segment.

In certain instances, the number of information bits may be less than the capacity of the transmission channels. In such instances, the available and unfilled bit positions may be filled with zero padding, by repeating some of the coded bits, or by some other scheme. The transmit power may also be reduced for some schemes.

Gray Mapping

In an embodiment, for each modulation scheme (e.g., QPSK, 16-QAM, 64-QAM, and so on) selected for use, the points in the signal constellation for the modulation scheme are defined using Gray mapping. The Gray mapping reduces the number of bit errors for more likely error events, as described in further detail below.

FIG. 5 is a diagram of a signal constellation for 16-QAM and a specific Gray mapping scheme. The signal constellation for 16-QAM includes 16 points, each of which is associated with a specific 4-bit value. For Gray mapping, the 4-bit values are associated with the points in the signal constellation such that the values for adjacent points (in the horizontal or vertical direction) differ by only one bit position. The values for points further way differ by more bit positions (e.g., the values for adjacent points in the diagonal direction differ by two bit positions).

Each group of four coded bits ($b_1 b_2 b_3 b_4$) is mapped to a specific point in the signal constellation associated with the same value as that of the four coded bits. For example, a value of ("0111") for the four coded bits is mapped to a point 512 in the signal constellation. This point then represents the modulation symbol for the four coded bits. For 16-QAM, each modulation symbol represents a specific one of the 16 points in the signal constellation, with the specific point being determined by the value of the four coded bits. Each modulation symbol can be expressed as a complex number ($c + jd$) and provided to the next processing element (i.e., MIMO processor 120 in FIG. 1).

At the receiver unit, the modulation symbols are received in the presence of noise and typically do not map to the exact location in the signal constellation. For the above example, the received modulation symbol for the transmitted coded bits ("0111") may not map to point 512 at the receiver unit. The noise may have caused the received modulation symbol to be mapped to another location in the signal constellation. Typically, there is greater likelihood of the received modulation symbol being mapped to a location near the correct location (e.g., near the points for "0101", "0011", "0110", or "1111"). Thus, the more likely error event is a received modulation symbol being

erroneously mapped to a point adjacent to the correct point. And since adjacent points in the signal constellation have values that differ by only one bit position, the Gray mapping reduces the number of error bits for more likely error events.

FIG. 5 shows a specific Gray mapping scheme for the 16-QAM signal constellation. Other Gray mapping schemes may also be used and are within the scope of the invention. The signal constellations for other modulation schemes (e.g., 8-PSK, 64-QAM, and so on) may also be mapped with similar or other Gray mapping schemes. For some modulation schemes such as 32-QAM and 128-QAM, a partial Gray mapping scheme may be used if a full Gray mapping scheme is not possible. Also, mapping schemes not based on Gray mapping may also be used and are within the scope of the invention.

MIMO Processing

FIG. 6 is a block diagram of an embodiment of a MIMO processor 120x, which is one implementation of MIMO processor 120 in FIG. 1. The modulation symbols may be transmitted on multiple frequency subchannels and possibly from multiple transmit antennas. When operating in the MIMO mode, the transmission on each frequency subchannel and from each transmit antenna represents non-duplicated data.

Within MIMO processor 120x, a demultiplexer (DEMUX) 610 receives and demultiplexes the modulation symbols into a number of subchannel symbol streams, S_1 through S_L, one subchannel symbol stream for each frequency subchannel used to transmit the symbols. Each subchannel symbol stream is then provided to a respective subchannel MIMO processor 612.

Each subchannel MIMO processor 612 may further demultiplex the received subchannel symbol stream into a number of (up to N_T) symbol sub-streams, one symbol sub-stream for each antenna used to transmit the modulation symbols. When the OFDM system is operated in the MIMO mode, each subchannel MIMO processors 612 pre-conditions the (up to) N_T modulation symbols in accordance with equation (1) described above to generate pre-conditioned modulation symbols, which are subsequently transmitted. In the MIMO mode, each pre-conditioned modulation symbol for a particular frequency subchannel of a particular transmit antenna represents a linear combination of (weighted) modulation symbols for up to N_T transmit antennas. Each of the (up to) N_T modulation symbols used to generate each pre-conditioned modulation symbol may be associated with a different signal constellation.

For each time slot, (up to) N_T pre-conditioned modulation symbols may be generated by each subchannel MIMO processor 612 and provided to (up to) N_T symbol combiners 616a through 616t. For example, subchannel MIMO processor 614a assigned to frequency subchannel 1 may provide up to N_T pre-conditioned modulation symbols for frequency subchannel 1 of antennas 1 through N_T. Similarly, subchannel MIMO processor 612*l* assigned to frequency subchannel L may provide up to N_T symbols for frequency subchannel L of antennas 1 through N_T. Each combiner 616 receives the pre-conditioned modulation symbols for the L frequency subchannels, combines the symbols for each time slot into a modulation symbol vector, V, and provides the modulation symbol vector to the next processing stage (i.e., modulator 122).

MIMO processor 120x thus receives and processes the modulation symbols to provide N_T modulation symbol vectors, V_1 through V_T, one modulation symbol vector for each transmit antenna. The collection of L pre-conditioned modulation symbols for each time slot of each antenna form a modulation symbol vector V of dimensionality L. Each element of the modulation symbol vector V is associated with a specific frequency subchannel having a unique subcarrier on which the modulation symbol is conveyed. The collection of the L modulation symbols are all orthogonal to one another. If not operating in a "pure" MIMO mode, some of the modulation symbol vectors may have duplicate information on specific frequency subchannels for different transmit antennas.

Subchannel MIMO processor 612 may be designed to provide the necessary processing to implement full channel state information (full-CSI) or partial-CSI processing for the MIMO mode. Full CSI includes sufficient characterization of the propagation path (i.e., amplitude and phase) between all pairs of transmit and receive antennas for each frequency subchannel. Partial CSI may include, for example, the SNR of the spatial subchannels. The CSI processing may be performed based on the available CSI information and on the selected frequency subchannels, transmit antennas, and so on. The CSI processing may also be enabled and disabled selectively and dynamically. For example, the CSI processing may be enabled for a particular data transmission and disabled for some other data transmissions. The CSI processing may be enabled under certain conditions, for example, when the communication link has adequate SNR. Full-CSI and partial-CSI processing is described in further detail in the aforementioned U.S Patent Application Serial No. 09/532,491.

FIG. 6 also shows an embodiment of modulator 122. The modulation symbol vectors V_1 through V_T from MIMO processor 120x are provided to

modulators 114a through 114t, respectively. In the embodiment shown in FIG. 6, each modulator 114 includes an IFFT 620, cycle prefix generator 622, and an upconverter 624.

IFFT 620 converts each received modulation symbol vector into its time-domain representation (which is referred to as an OFDM symbol) using the inverse fast Fourier transform (IFFT). IFFT 620 can be designed to perform the IFFT on any number of frequency subchannels (e.g., 8, 16, 32, and so on). In an embodiment, for each modulation symbol vector converted to an OFDM symbol, cycle prefix generator 622 repeats a portion of the time-domain representation of the OFDM symbol to form a transmission symbol for the specific antenna. The cyclic prefix insures that the transmission symbol retains its orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious path effects. The implementation of IFFT 620 and cycle prefix generator 622 is known in the art and not described in detail herein.

The time-domain representations from each cycle prefix generator 622 (i.e., the "transmission" symbols for each antenna) are then processed by upconverter 624, converted into an analog signal, modulated to a RF frequency, and conditioned (e.g., amplified and filtered) to generate an RF modulated signal, which is then transmitted from the respective antenna 124.

OFDM modulation is described in further detail in a paper entitled "Multicarrier Modulation for Data Transmission : An Idea Whose Time Has Come," by John A.C. Bingham, IEEE Communications Magazine, May 1990, which is incorporated herein by reference.

For an OFDM system not operated in the MIMO mode, MIMO processor 120 may be removed or disabled and the modulation symbols may be grouped into the modulation symbol vector V without any pre-conditioning. This vector is then provided to modulator 122. And for an OFDM system operated with transmit diversity (and not in the MIMO mode), demultiplexer 614 may be removed or disabled and the (same) pre-conditioned modulation symbols are provided to (up to) N_T combiners.

As shown in FIG. 2, a number of different transmissions (e.g., voice, signaling, data, pilot, and so on) may be transmitted by the system. Each of these transmissions may require different processing.

FIG. 7 is a block diagram of an embodiment of a system 110y capable of providing different processing for different transmissions. The aggregate input data, which includes all information bits to be transmitted by system 110y, is provided to a demultiplexer 710. Demultiplexer 710 demultiplexes the input

data into a number of (K) channel data streams, B_1 through B_k. Each channel data stream may correspond to, for example, a signaling channel, a broadcast channel, a voice call, or a traffic data transmission. Each channel data stream is provided to a respective encoder/channel interleaver/puncturer/symbol mapping element 712 that encodes the data using a particular encoding scheme selected for that channel data stream, interleaves the encoded data based on a particular interleaving scheme, punctures the interleaved code bits, and maps the interleaved data into modulation symbols for the one or more transmission channels used for transmitting that channel data stream.

The encoding can be performed on a per channel basis (i.e., on each channel data stream, as shown in FIG. 7). However, the encoding may also be performed on the aggregate input data (as shown in FIG. 1), on a number of channel data streams, on a portion of a channel data stream, across a set of frequency subchannels, across a set of spatial subchannels, across a set of frequency subchannels and spatial subchannels, across each frequency subchannel, on each modulation symbol, or on some other unit of time, space, and frequency.

The modulation symbol stream from each encoder/channel interleaver/puncturer/symbol mapping element 712 may be transmitted on one or more frequency subchannels and via one or more spatial subchannels of each frequency subchannel. A MIMO processor 120y receives the modulation symbol streams from elements 712. Depending on the mode to be used for each modulation symbol stream, MIMO processor 120y may demultiplex the modulation symbol stream into a number of subchannel symbol streams. In the embodiment shown in FIG. 7, modulation symbol stream S_1 is transmitted on one frequency subchannel and modulation symbol stream S_K is transmitted on L frequency subchannels. The modulation stream for each frequency subchannel is processed by a respective subchannel MIMO processor, demultiplexed, and combined in similar manner as that described in FIG. 6 to form a modulation symbol vector for each transmit antenna.

In general, the transmitter unit codes and modulates data for each transmission channel based on information descriptive of the channel's transmission capability. This information is typically in the form of partial-CSI or full-CSI described above. The partial or full-CSI for the transmission channels to be used for a data transmission is typically determined at the receiver unit and reported back to the transmitter unit, which then uses the information to code and modulate data accordingly. The techniques described herein are applicable for multiple parallel transmission channels supported by

MIMO, OFDM, or any other communication scheme (e.g., a CDMA scheme) capable of supporting multiple parallel transmission channels.

Demodulation and Decoding

FIG. 8 is a block diagram of an embodiment of a decoding portion of system 150. For this embodiment, a Turbo encoder is used to encode the data prior to transmission. A Turbo decoder is correspondingly used to decode the received modulation symbols.

As shown in FIG. 8, the received modulation symbols are provided to a bit log-likelihood ratio (LLR) calculation unit 158x, which calculates the LLRs of the bits that make up each modulation symbol. Since a Turbo decoder operates on LLRs (as oppose to bits), bit LLR calculation unit 158x provides an LLR for each received coded bit. The LLR for each received coded bit is the logarithm of the probability that the received coded bit is a zero divided by the probability that the received coded bit is a one.

As described above, M coded bits (b_1, b_2, ... b_M) are grouped to form a single non-binary symbol S, which is then mapped to a modulation symbol T(S) (i.e., modulated to a high-order signal constellation). The modulation symbol is processed, transmitted, received, and further processed to provide a received modulation symbol R(S). The LLR of coded bit b_m in the received modulation symbol can be computed as:

$$\begin{aligned} LLR(b_m) &= \log\left(\frac{P(b_m=0)}{P(b_m=1)}\right) \\ &= \log\big(P\big(R(S) \mid b_m=0\big)\big) - \log\big(P\big(R(S) \mid b_m=1\big)\big) \\ &= \log\left(\sum_{T(S):b_m=0} P\big(R(S) \mid T(S)\big)\right) - \log\left(\sum_{T(S):b_m=1} P\big(R(S) \mid T(S)\big)\right) \end{aligned} \qquad \text{Eq (2)}$$

where $P\big(R(S) \mid b_m=0\big)$ is the probability of bit b_m being a zero based on the received symbol R(S). Approximations may also be used in computing the LLRs.

De-puncturer 159 then inserts "erasures" for code bits that have been deleted (i.e., punctured) at the transmitter. The erasures typically have a value of zero ("0"), which is indicative of the punctured bit being equally likely to be a zero or a one.

From equation (2), it can be noted that the LLRs for the received coded bits within a modulation symbol tend to be correlated. This correlation can be

broken up by interleaving the coded bits prior to modulation. As shown in FIG. 1, the channel interleaving advantageously performs the decorrelation of the coded bits in each modulation symbol.

The coded bit LLRs are provided to a channel deinterleaver 160 and deinterleaved in a manner complementary to the channel interleaving performed at the transmitter. The channel deinterleaved LLRs corresponding to the received information, tail, and parity bits are then provided to a Turbo decoder 162x.

Turbo decoder 162x includes summers 810a and 810b, decoders 812a and 812b, a code interleaver 814, a code deinterleaver 816, and a detector 818. In an embodiment, each decoder 812 is implemented as a soft-input/soft-output (SISO) maximum a posterior (MAP) decoder.

Summer 810a receives and sums the LLRs of the received information bits, LLR(x'), and the extrinsic information from deinterleaver 816 (which is set to zeros on the first iteration), and provides refined LLRs. The refined LLRs are associated with greater confidence in the detected values of the received information bits.

Decoder 812a receives the refined LLRs from summer 810a and the LLRs of the received tail and parity bits from the first constituent encoder, LLR(y'), and decodes the received LLRs to generate extrinsic information indicative of corrections in the probability values for the received information bits. The extrinsic information from decoder 812a are summed with the received information bit LLRs by summer 810b, and the refined LLRs are stored to code interleaver 814. Code interleaver 814 implements the same code interleaving used at the Turbo encoder (e.g., the same as code interleaver 314 in FIG. 3B).

Decoder 812b receives the interleaved LLRs from interleaver 814 and the LLRs of the received tail and parity bits from the second constituent encoder, LLR(z'), and decodes the received LLRs to generate extrinsic information indicative of further corrections in the probability values for the received information bits. The extrinsic information from decoder 812b is stored to code deinterleaver 816, which implements a deinterleaving scheme complementary to the interleaving scheme used for interleaver 814.

The decoding of the received coded bit LLRs is iterated a number of times. With each iteration, greater confidence is gained for the refined LLRs. After all the decoding iterations have been completed, the final refined LLRs are provided to detector 818, which provides values for the received information bits based on the LLRs.

Other types of decoder may also be used beside the SISO MAP decoder such as one that implements the soft output Viterbi algorithm (SOVA). The design of the decoder is typically dependent on the particular Turbo coding scheme used at the transmitter.

Turbo decoding is described in greater detail by Steven S. Pietrobon in a paper entitled "Implementation and Performance of a Turbo/Map Decoder," International Journal of Satellite Communications, Vol. 16, 1998, pp. 23-46, which is incorporated herein by reference.

Modulation Scheme and Coding Rate

The achieved SNR of each transmission channel supports a particular number of information bits per modulation symbol (i.e., a particular information bit rate) for a desired level of performance (e.g., 1% FER). This information bit rate may be supported by a number of different modulation schemes. For example, a bit rate of 1.5 information bits/modulation symbol may be supported by QPSK, 8-PSK, 16-QAM, or any higher order modulation scheme. Each modulation scheme is able to transmit a particular number of coded bits per modulation symbol.

Depending on the selected modulation scheme, a corresponding coding rate is selected such that the required number of coded bits is provided for the number of information bits for each modulation symbol. For the above example, QPSK, 8-PSK, and 16-QAM are respectively able to transmit 2, 3, and 4 coded bits per modulation symbol. For an information bit rate of 1.5 information bits/modulation symbol, coding rates of 3/4, 1/2, and 3/8 are used to generate the required number of coded bits for QPSK, 8-PSK, and 16-QAM, respectively. Thus, different combinations of modulation scheme and coding rate may be used to support a particular information bit rate.

In certain embodiments of the invention, a "weak" binary code (i.e., a high coding rate) is used in conjunction with a low-order modulation scheme for the supported bit rate. Through a series of simulation, it is observed that a lower order modulation scheme in combination with a weaker code may offer better performance than a higher order modulation scheme with a stronger code. This result may be explained as follows. The LLR decoding metrics of binary Turbo codes in an AWGN channel is near optimal for the Turbo decoding algorithm. However, for the Gray mapped high-order modulation scheme, the optimal LLR metrics are generated for each received modulation symbol and not each received bit. The symbol LLR metrics are then broken to yield bit LLR metrics for the binary code decoder. Some information is lost

during the break-up process, and using the bit decoding metrics may result in non-optimal performance. The lower order modulation schemes correspond to fewer bits per symbol, which may experience less of the break-up loss and therefore provide better performance than the higher order modulation scheme counterparts.

In accordance with an aspect of the invention, in order to achieve certain spectrum efficiency, a code with a coding rate of between, and inclusive of, $n/(n+1)$ to $n/(n+2)$ is used with an appropriate modulation scheme, where n is the number of information bits per modulation symbol. This coding rate may be easily achieved with a fixed code (e.g., the rate 1/3 Turbo code described above) in combination with a variable puncturing scheme. To achieve a high coding rate, the tail and parity bits may be heavily punctured and the unpunctured tail and parity bits may be evenly distributed over the information bits.

Framing

For many communication systems, it is convenient to define data packets (i.e., logical frames) with fixed sizes. For example, a system may define three different packets having sizes of 1024, 2048, and 4096 bits. These defined data packets simplify some of the processing at both the transmitter and receiver.

For an OFDM system, a physical frame may be defined to include (1) an integer number of OFDM symbols, (2) a particular number of modulation symbols on one or more transmission channels, (3) or some other units. As described above, because of the time-variant nature of the communication link, the SNR of the transmission channels may vary over time. Consequently, the number of information bits which may be transmitted on each time slot for each transmission channel will likely vary over time, and the number of information bits in each physical frame will also likely vary over time.

In one embodiment, a logical frame is defined such that it is independent of the OFDM symbols. In this embodiment, the information bits for each logical frame are encoded/punctured, and the coded bits for the logical frame are grouped and mapped to modulation symbols. In one simple implementation, the transmission channels are sequentially numbered. The coded bits are then used to form as many modulation symbols as needed, in the sequential order of the transmission channels. A logical frame (i.e., a data packet) may be defined to start and end at modulation symbol boundaries. In this implementation, the logical frame may span more than one OFDM symbol

and may further cross OFDM symbol boundaries. Moreover, each OFDM symbol may include coded bits from multiple data packets.

In another embodiment, a logical frame is defined based on a physical unit. For example, a logical frame may be defined to include (1) a number of modulation symbols on one or more transmission channels, (2) one or more OFDM symbols, or (3) a number of modulation symbols defined in some other manner.

The use of punctured binary Turbo code and Gray mapping (BTC-GM) for high-order modulation provides numerous advantages. The BTC-GM scheme is simpler to implement than the more optimal but more complicated Turbo trellis coded modulation (TTCM) scheme, yet can achieve performance close to that of TTCM. The BTC-GM scheme also provides a high degree of flexibility because of the ease of implementing different coding rate by simply adjusting the variable puncturing. The BTC-GM scheme also provides robust performance under different puncturing parameters. Also, currently available binary Turbo decoders may be used, which may simply the implementation of the receiver. However, in certain embodiments, other coding schemes may also be used and are within the scope of the invention.

The foregoing description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS

1. In a wireless communication system, a method for preparing data for transmission on a plurality of transmission channels, wherein each transmission channel is operative to transmit a respective sequence of modulation symbols, the method comprising:

determining a number of information bits per modulation symbol supported by each transmission channel;

identifying a modulation scheme for each transmission channel such that the determined number of information bits per modulation symbol is supported;

determining a coding rate for each transmission channel based at least on the determined number of information bits per modulation symbol and the identified modulation scheme for the transmission channel, wherein at least two transmission channels are associated with different coding rates;

encoding a plurality of information bits in accordance with a particular encoding scheme to provide a plurality of coded bits;

puncturing the plurality of coded bits in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the plurality of transmission channels; and

adjusting the puncturing to achieve the different coding rates for the at least two transmission channels.

2. The method of claim 1, wherein the wireless communication system is a multiple-input multiple-output (MIMO) system with a plurality of transmit antennas and a plurality of receive antennas.

3. The method of claim 1, wherein the wireless communication system is an orthogonal frequency division modulation (OFDM) communication system.

4. The method of claim 3, wherein the OFDM communication system is operated as a multiple-input multiple-output (MIMO) system with a plurality of transmit antennas and a plurality of receive antennas.

5. The method of claim 4, wherein the OFDM system is operative to transmit data on a plurality of frequency subchannels, and wherein each transmission channel corresponds to a spatial subchannel of a frequency subchannel in the OFDM system.

6. The method of claim 1, wherein the puncturing is based on transmission capabilities of the plurality of transmission channels.

7. The method of claim 6, wherein the transmission capabilities are determined from channel state information (CSI) derived for the plurality of transmission channels.

8. The method of claim 7, wherein the CSI includes signal-to-noise ratio (SNR) information for the plurality of transmission channels.

9. The method of claim 7, wherein the CSI includes information related to transmission characteristics from transmit antennas to the receive antennas.

10. The method of claim 7, wherein the CSI includes eigenmode information related to transmission characteristics from transmit antennas to the receive antennas.

11. The method of claim 6, further comprising:

grouping transmission channels having similar transmission capabilities to segments, and

wherein the puncturing is performed for each segment.

12. The method of claim 11, further comprising:

assigning a group of coded bits to each segment, and

wherein the puncturing is performed on the group of coded bits assigned to each segment.

13. The method of claim 11, wherein each segment includes transmission channels having SNR within a particular SNR range.

14. The method of claim 1, wherein the encoding is achieved via a Turbo code.

15. The method of claim 14, wherein the encoding provides a plurality of tail and parity bits for the plurality of information bits, and wherein the puncturing is performed on the plurality of tail and parity bits.

16. The method of claim 14, wherein the puncturing is performed such that unpunctured tail and parity bits are approximately evenly distributed over the plurality of information bits.

17. The method of claim 14, wherein the Turbo code includes two constituent codes operative to provide two streams of tail and parity bits, and wherein the puncturing is performed such that approximately equal number of tail and parity bits are deleted from the two streams of tail and parity bits.

18. The method of claim 1, wherein the coding rate for each transmission channel is selected to be between, and inclusive of, n/(n+1) and n/(n+2), where n is the number of information bits per modulation symbol supported by the transmission channel.

19. The method of claim 1, wherein the coding rate for each transmission channel is 1/2 or higher.

20. The method of claim 1, wherein the encoding is achieved via a convolutional code.

21. The method of claim 1, wherein the encoding is achieved via a block code.

22. The method of claim 1, further comprising:

inserting padding bits to fill available but unfilled bit positions in the plurality of transmission channels.

23. The method of claim 1, further comprising:

repeating at least some of the coded bits to fill available but unfilled bit positions in the plurality of transmission channels.

24. The method of claim 1, further comprising:

interleaving the plurality of coded bits.

25. The method of claim 24, wherein the puncturing is performed on interleaved coded bits.

26. The method of claim 24, wherein the encoding is achieved via a Turbo code comprised of two constituent codes, and wherein the plurality of information bits, a plurality of tail and parity bits from a first constituent code, and a plurality of tail and parity bits from a second constituent code are separately interleaved.

27. The method of claim 1, further comprising:

forming non-binary symbols for the plurality of transmission channels, wherein each non-binary symbol includes a group of unpunctured coded bits; and

mapping each non-binary symbol to a respective modulation symbol.

28. The method of claim 27, further comprising:

interleaving the plurality of coded bits, and

wherein the non-binary symbols are formed from the interleaved coded bits.

29. The method of claim 27, wherein the modulation scheme for each transmission channel is associated with a respective signal constellation having a plurality of points, and wherein each modulation symbol is representative of a particular point in the signal constellation for the modulation scheme.

30. The method of claim 29, wherein the plurality of points in each signal constellation are assigned with values based on a particular Gray mapping scheme.

31. The method of claim 30, wherein the values are assigned to the plurality of points in each signal constellation such that values for adjacent points in the signal constellation differ by one bit position.

32. The method of claim 1, further comprising:

adapting to changes in the plurality of transmission channels by repeating the determining the number of information bits per modulation symbol, the identifying the modulation scheme, and the determining the coding rate.

33. The method of claim 1, wherein the modulation scheme for each transmission channel supports transmission of two or more coded bits per modulation symbol.

34. The method of claim 1, wherein the transmission on the plurality of transmission channels are intended for a single recipient receiving device.

35. In an orthogonal frequency division modulation (OFDM) communication system, a method for preparing data for transmission on a plurality of transmission channels, wherein each transmission channel is operative to transmit a respective sequence of modulation symbols, the method comprising:

determining a number of information bits per modulation symbol supported by each transmission channel;

identifying a modulation scheme for each transmission channel such that the determined number of information bits per modulation symbol is supported;

determining a coding rate for each transmission channel based at least on the determined number of information bits per modulation symbol and the identified modulation scheme for the transmission channel, wherein at least two transmission channels are associated with different coding rates;

encoding a plurality of information bits in accordance with a particular Turbo code to provide a plurality of tail and parity bits;

interleaving the plurality of information and tail and parity bits in accordance with a particular interleaving scheme;

puncturing the plurality of interleaved bits in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the plurality of transmission channels, wherein the puncturing is adjusted to achieve the different coding rates for the at least two transmission channels;

forming non-binary symbols for the plurality of transmission channels, wherein each non-binary symbol includes a group of unpunctured coded bits; and

mapping each non-binary symbol to a respective modulation symbol.

36. A wireless communication system operative to transmit data on a plurality of transmission channels, wherein each transmission channel is used to transmit a respective sequence of modulation symbols, the system comprising:

an encoder configured to encode a plurality of information bits in accordance with a particular encoding scheme to provide a plurality of coded bits, and to puncture the plurality of coded bits in accordance with a particular puncturing scheme to provide a number of unpunctured coded bits for the plurality of transmission channels, wherein each transmission channel is capable of transmitting a particular number of information bits per modulation symbol via a particular modulation scheme selected for the transmission channel, wherein each transmission channel is further associated with a particular coding rate based at least on the number of information bits per modulation symbol supported by the transmission channel and its modulation scheme, wherein at least two transmission channels are associated with different coding rates, and wherein the encoder is further configured to adjust the puncturing to achieve the different coding rates for the at least two transmission channels.

37. The system of claim 36, further comprising:

a channel interleaver coupled to the encoder and configured to interleave the plurality of coded bits, and

wherein the encoder is configured to puncture the interleaved bits.

38. The system of claim 37, further comprising:

a symbol mapping element coupled to the channel interleaver and configured to form non-binary symbols for the plurality of transmission channels, and to map each non-binary symbol to a respective modulation symbol, wherein each non-binary symbol includes a group of unpunctured coded bits.

39. The system of claim 38, further comprising:

a signal processor coupled to the symbol mapping element and configured to pre-condition the modulation symbols for the plurality of transmission channels to implement a multiple-input multiple-output (MIMO) transmission.

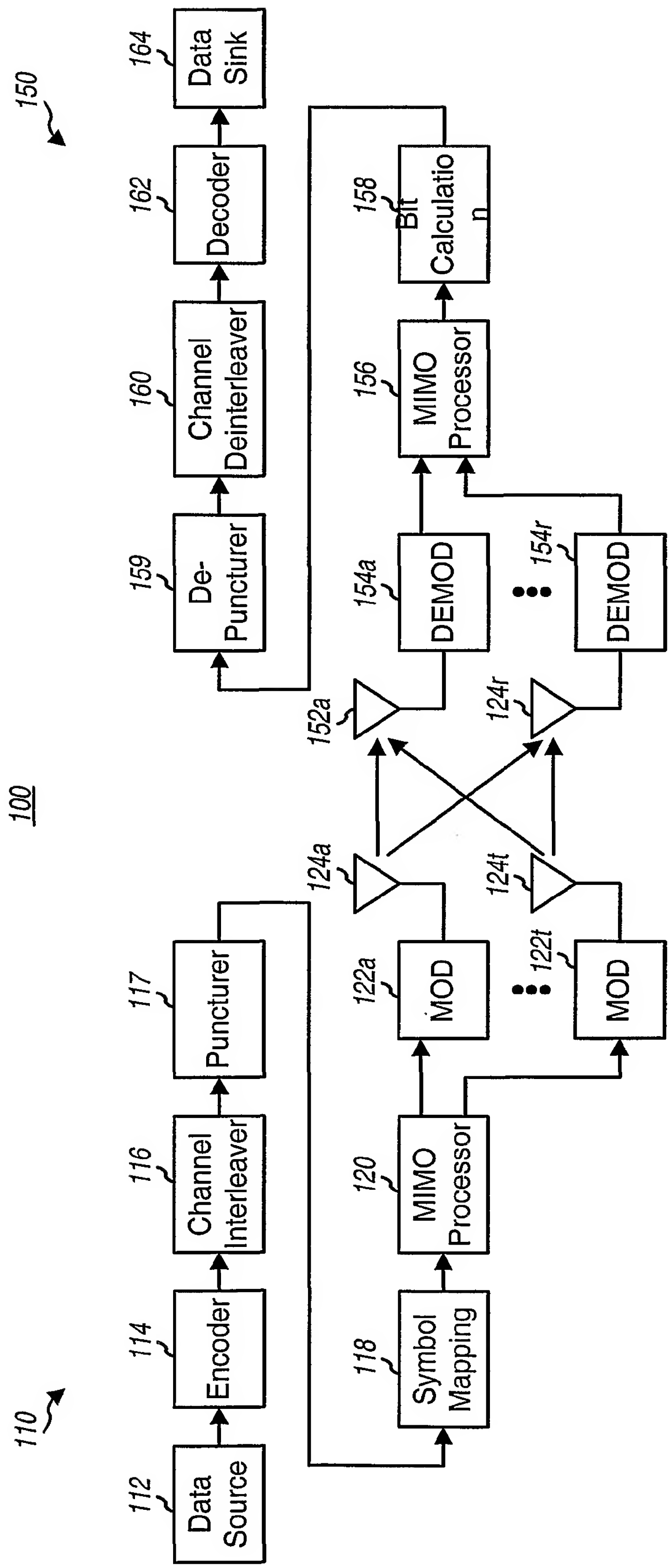

100
110
150
Data Source
112
Encoder
114
Channel Interleaver
116
Puncturer
117
Symbol Mapping
118
MIMO Processor
120
MOD
122a
MOD
122t
124a
124t
152a
124r
DEMOD
154a
DEMOD
154r
MIMO Processor
156
Bit Calculation
158
De-Puncturer
159
Channel Deinterleaver
160
Decoder
162
Data Sink
164

FIG. 1

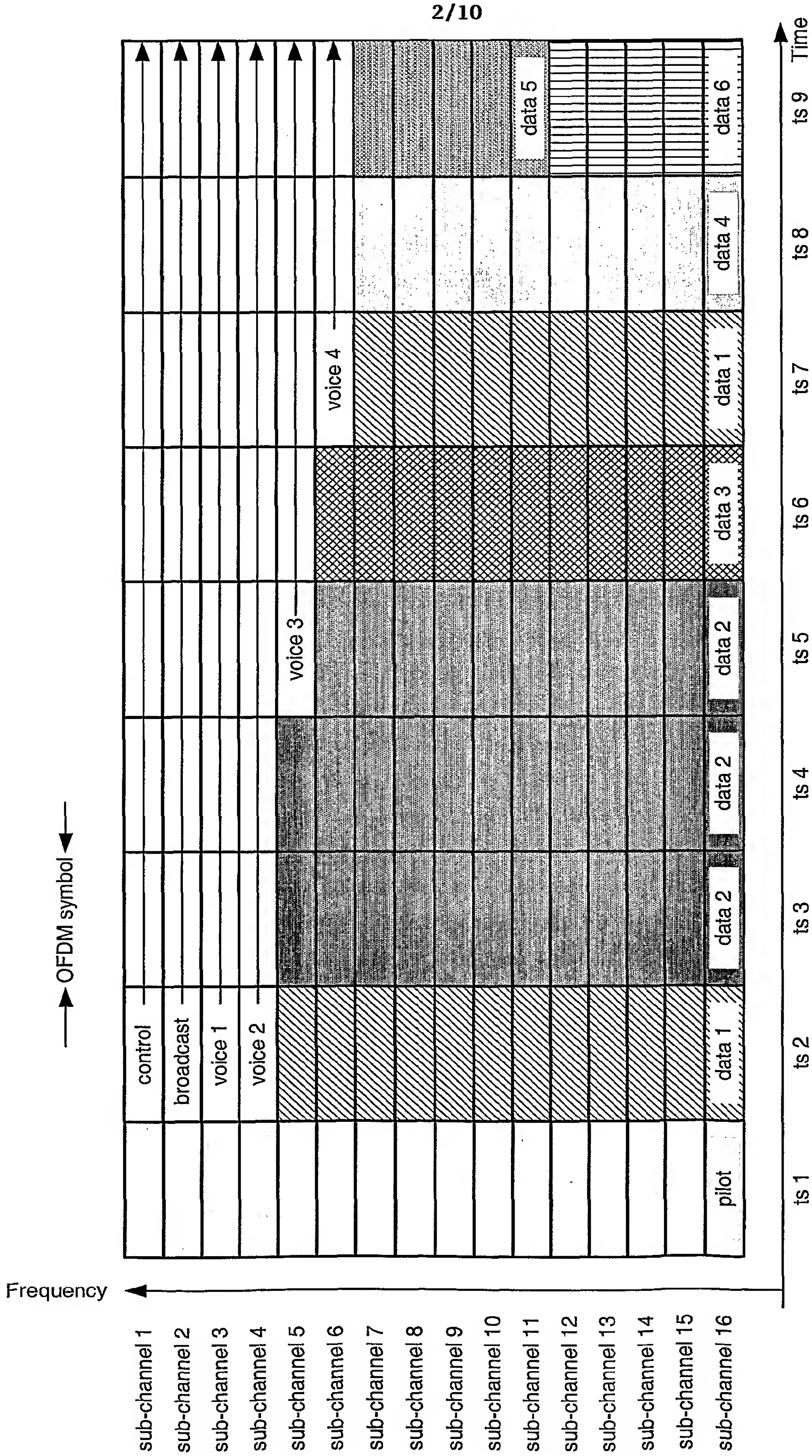

OFDM symbol
Frequency
Time
sub-channel 1
sub-channel 2
sub-channel 3
sub-channel 4
sub-channel 5
sub-channel 6
sub-channel 7
sub-channel 8
sub-channel 9
sub-channel 10
sub-channel 11
sub-channel 12
sub-channel 13
sub-channel 14
sub-channel 15
sub-channel 16
pilot
control
broadcast
voice 1
voice 2
voice 3
voice 4
data 1
data 2
data 2
data 2
data 3
data 1
data 4
data 5
data 6
ts 1
ts 2
ts 3
ts 4
ts 5
ts 6
ts 7
ts 8
ts 9

FIG. 2

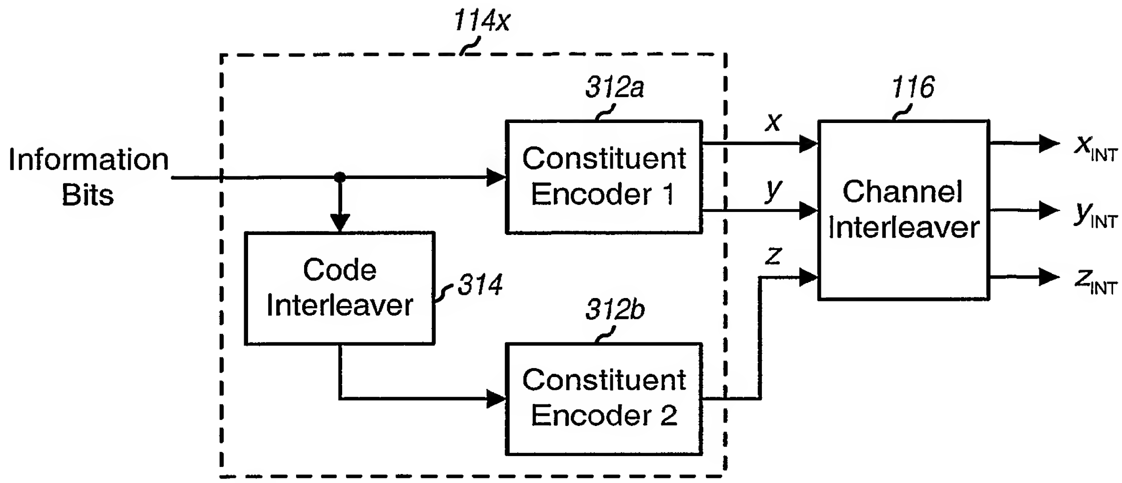

114x
312a
116
Information Bits
Constituent Encoder 1
x
y
z
Channel Interleaver
x_{INT}
y_{INT}
z_{INT}
Code Interleaver
314
312b
Constituent Encoder 2

FIG. 3A

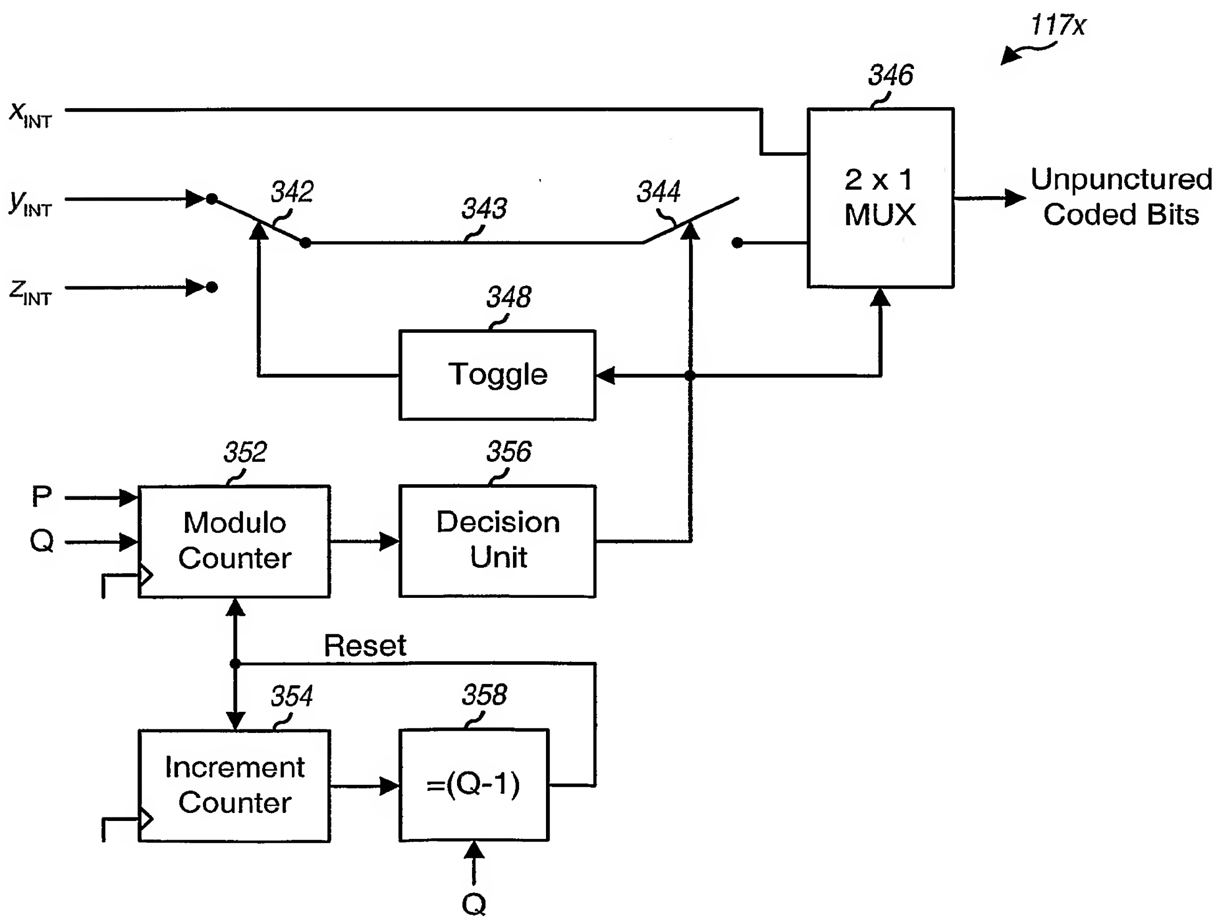
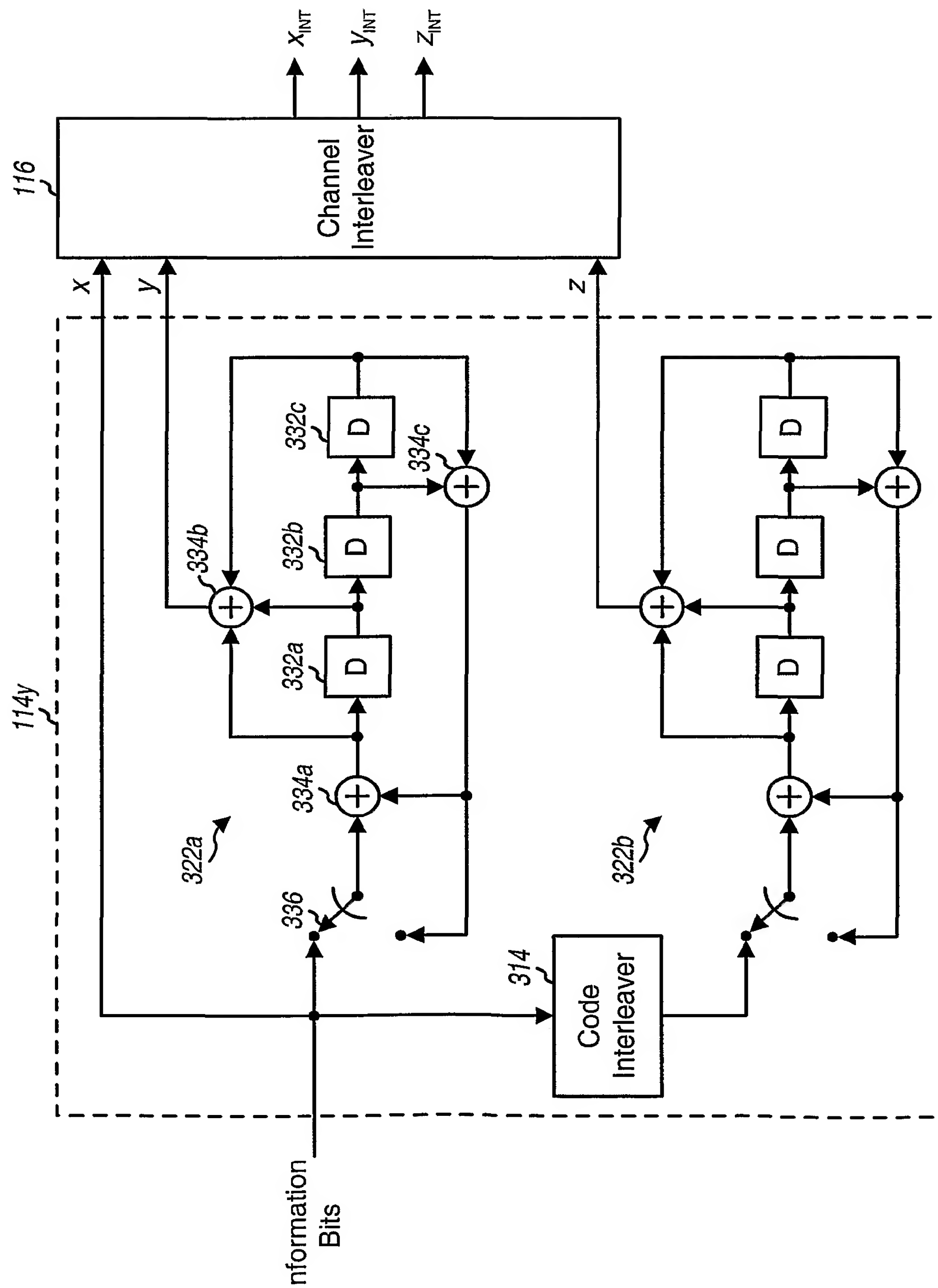

117x
346
x_{INT}
y_{INT}
z_{INT}
342
343
344
2 x 1 MUX
Unpunctured Coded Bits
348
Toggle
352
356
P
Q
Modulo Counter
Decision Unit
Reset
354
358
Increment Counter
=(Q-1)
Q

FIG. 3C



116
Channel Interleaver
x_{INT}
y_{INT}
z_{INT}
x
y
z
114y
332a
332b
332c
334a
334b
334c
D
322a
322b
336
314
Code Interleaver
Information Bits

FIG. 3B

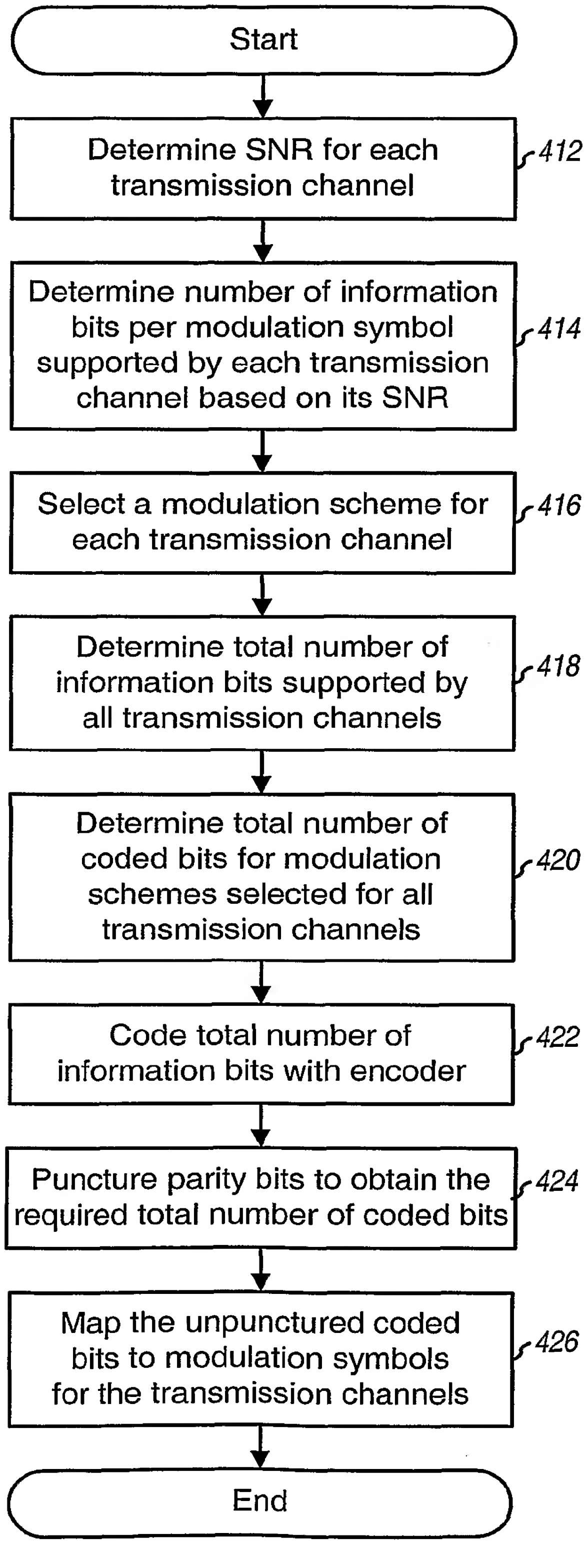

Start
Determine SNR for each transmission channel
412
Determine number of information bits per modulation symbol supported by each transmission channel based on its SNR
414
Select a modulation scheme for each transmission channel
416
Determine total number of information bits supported by all transmission channels
418
Determine total number of coded bits for modulation schemes selected for all transmission channels
420
Code total number of information bits with encoder
422
Puncture parity bits to obtain the required total number of coded bits
424
Map the unpunctured coded bits to modulation symbols for the transmission channels
426
End

FIG. 4A

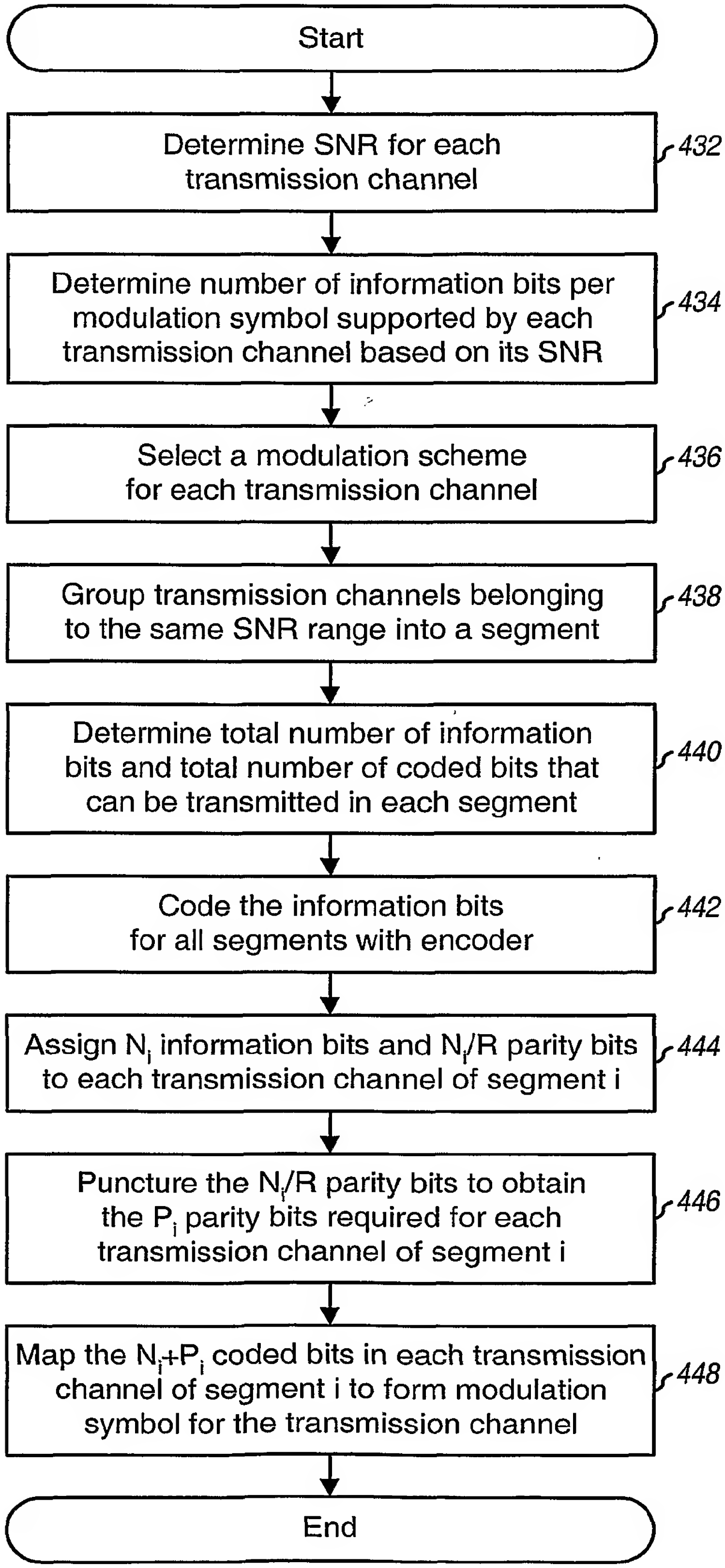

Start
Determine SNR for each transmission channel
432
Determine number of information bits per modulation symbol supported by each transmission channel based on its SNR
434
Select a modulation scheme for each transmission channel
436
Group transmission channels belonging to the same SNR range into a segment
438
Determine total number of information bits and total number of coded bits that can be transmitted in each segment
440
Code the information bits for all segments with encoder
442
Assign N_i information bits and N_i/R parity bits to each transmission channel of segment i
444
Puncture the N_i/R parity bits to obtain the P_i parity bits required for each transmission channel of segment i
446
Map the N_i+P_i coded bits in each transmission channel of segment i to form modulation symbol for the transmission channel
448
End

FIG. 4B

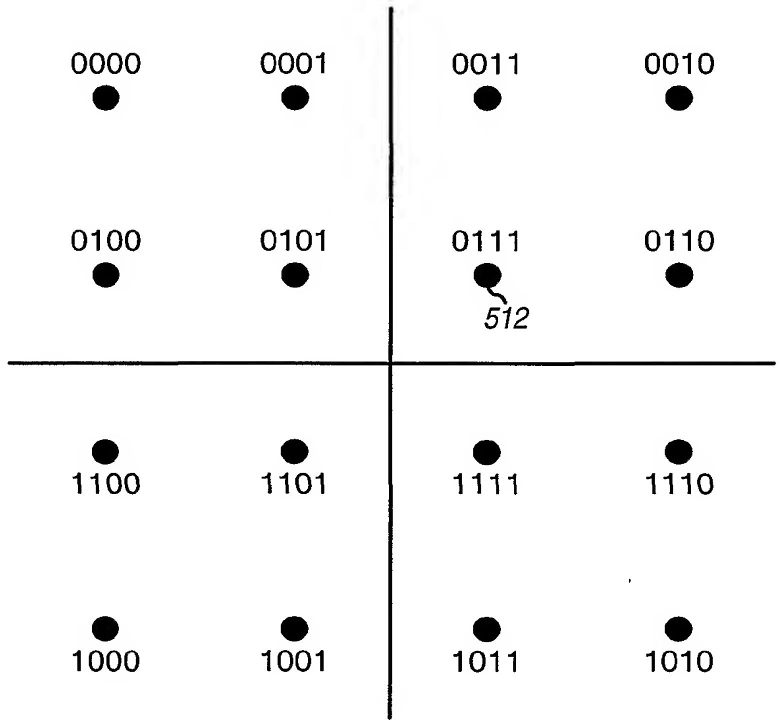

0000
0001
0011
0010
0100
0101
0111
0110
512
1100
1101
1111
1110
1000
1001
1011
1010

FIG. 5

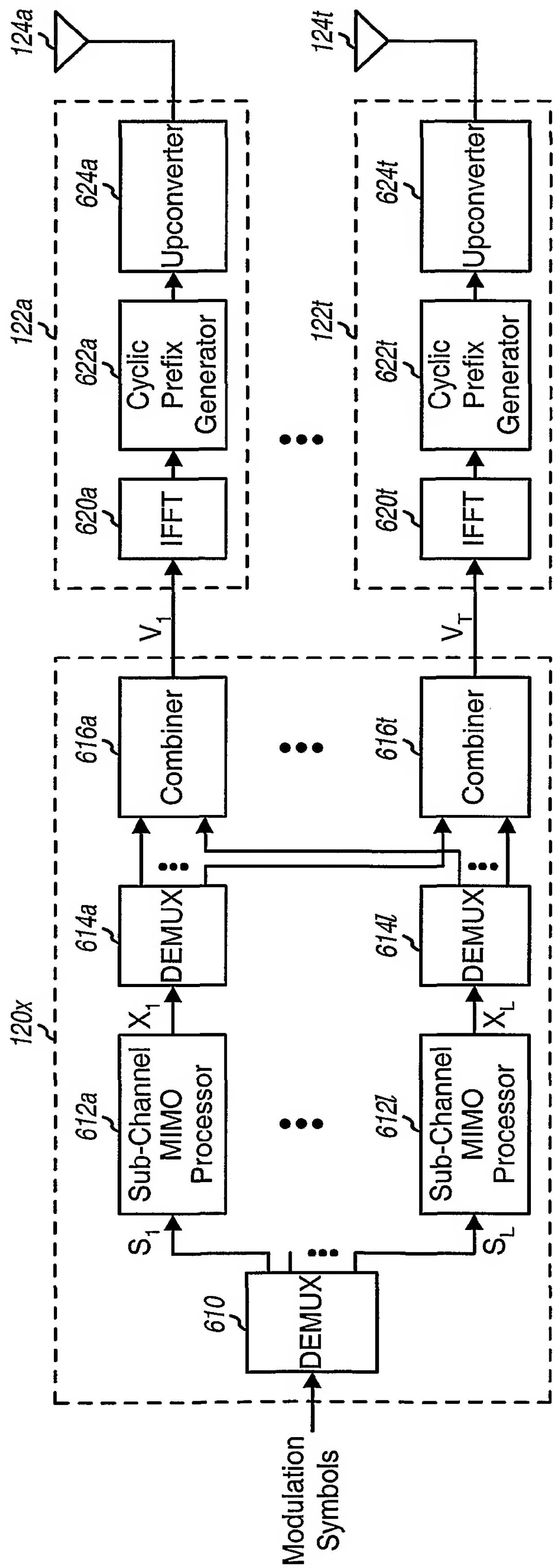

Modulation Symbols
610
DEMUX
120x
S_1
S_L
612a
Sub-Channel MIMO Processor
612l
Sub-Channel MIMO Processor
X_1
X_L
614a
DEMUX
614l
DEMUX
616a
Combiner
616t
Combiner
V_1
V_T
122a
620a
IFFT
622a
Cyclic Prefix Generator
624a
Upconverter
124a
122t
620t
IFFT
622t
Cyclic Prefix Generator
624t
Upconverter
124t

FIG. 6

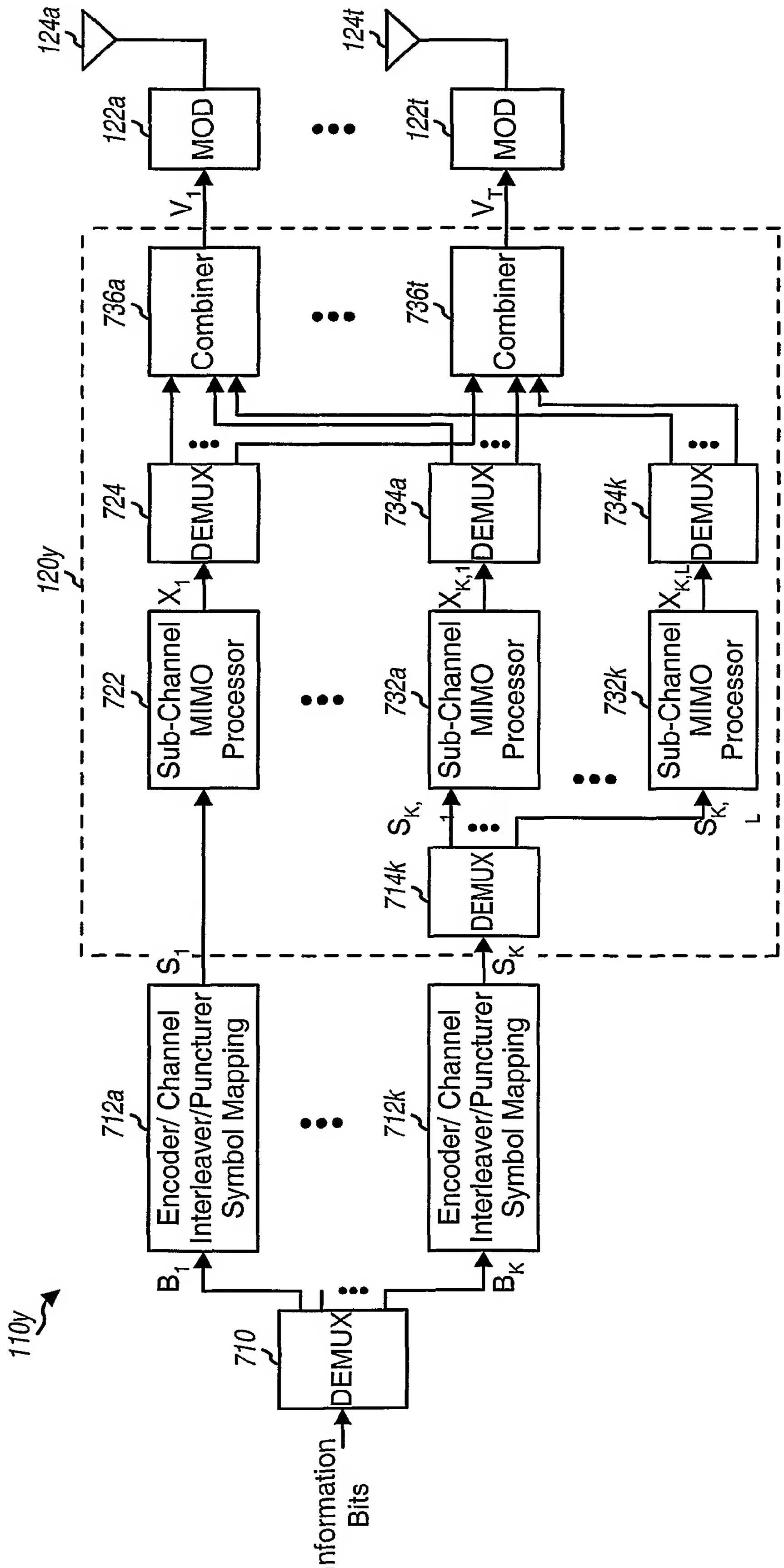

110y
Information Bits
710
DEMUX
B1
BK
712a
Encoder/ Channel Interleaver/Puncturer Symbol Mapping
712k
Encoder/ Channel Interleaver/Puncturer Symbol Mapping
S1
SK
120y
714k
DEMUX
SK,1
SK,L
722
Sub-Channel MIMO Processor
732a
Sub-Channel MIMO Processor
732k
Sub-Channel MIMO Processor
X1
XK,1
XK,L
724
DEMUX
734a
DEMUX
734k
DEMUX
736a
Combiner
736t
Combiner
V1
VT
122a
MOD
122t
MOD
124a
124t

FIG. 7

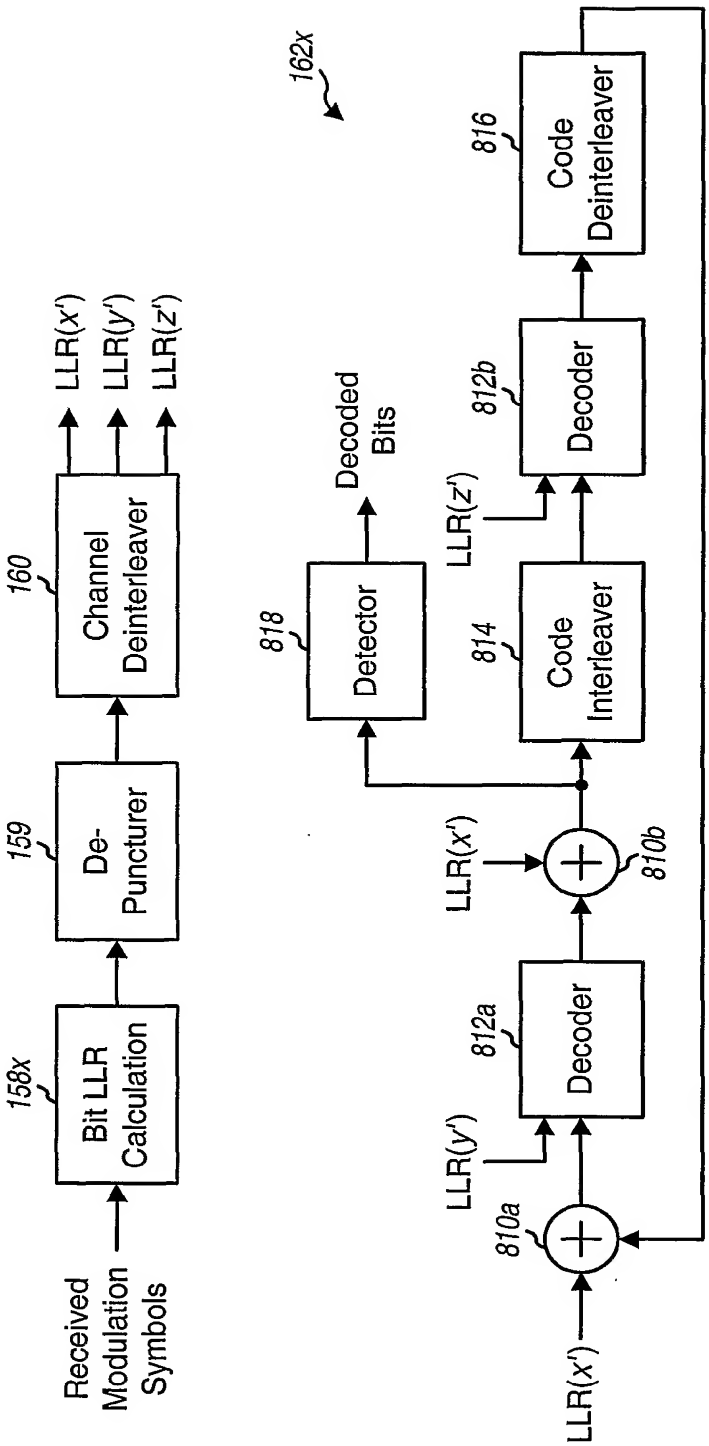

162x
Received Modulation Symbols
158x
Bit LLR Calculation
159
De-Puncturer
160
Channel Deinterleaver
LLR(x')
LLR(y')
LLR(z')
LLR(x')
810a
LLR(y')
812a
Decoder
LLR(x')
810b
818
Detector
Decoded Bits
814
Code Interleaver
LLR(z')
812b
Decoder
816
Code Deinterleaver

FIG. 8

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/02143

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L1/00 H04L27/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, COMPENDEX, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X A	US 5 197 061 A (LE FLOCH BERNARD ET AL) 23 March 1993 (1993-03-23) column 2, line 67 -column 3, line 10 column 3, line 15 - line 18 column 4, line 29 - line 40 column 4, line 59 - line 62 column 5, line 18 - line 21 column 5, line 54 - line 68 column 6, line 40 - line 47 column 9, line 51 - line 59 --- -/--	1,3,6, 35,36 2,4,5, 7-34, 37-39

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

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Date of the actual completion of the international search	Date of mailing of the international search report
3 July 2002	23/07/2002

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Schiffer, A

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

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A	MATSUOKA H ET AL: "Adaptive modulation system with variable coding rate concatenated code for high quality multi-media communication systems" VEHICULAR TECHNOLOGY CONFERENCE, 1996. MOBILE TECHNOLOGY FOR THE HUMAN RACE., IEEE 46TH ATLANTA, GA, USA 28 APRIL-1 MAY 1996, NEW YORK, NY, USA,IEEE, US, 28 April 1996 (1996-04-28), pages 487-491, XP010162435 ISBN: 0-7803-3157-5	1,6-8, 30,35,36
A	abstract page 487, right-hand column, paragraph 1 page 487, right-hand column, paragraph 3 page 488, left-hand column, paragraph 1 page 488, left-hand column, paragraph 3 - paragraph 4 page 488, right-hand column, paragraph 5 ---	2-5, 9-29, 31-34, 37-39
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INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 02/02143

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT		
Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
A	SAMPEI S ET AL: "ADAPTIVE MODULATION/TMDA SCHEME FOR LARGE CAPACITY PERSONAL MULTI-MEDIA COMMUNICATION SYSTEMS" IEICE TRANSACTIONS ON COMMUNICATIONS, INSTITUTE OF ELECTRONICS INFORMATION AND COMM. ENG. TOKYO, JP, vol. E77-B, no. 9, 1 September 1994 (1994-09-01), pages 1096-1103, XP000474107 ISSN: 0916-8516 page 1096, right-hand column, paragraph 3 page 1097, left-hand column, paragraph 3 page 1097, right-hand column, paragraph 5 page 1098, right-hand column, paragraph 6 page 1100, left-hand column, paragraph 2	1-39

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/US 02/02143

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			EP	1123613	A1	16-08-2001
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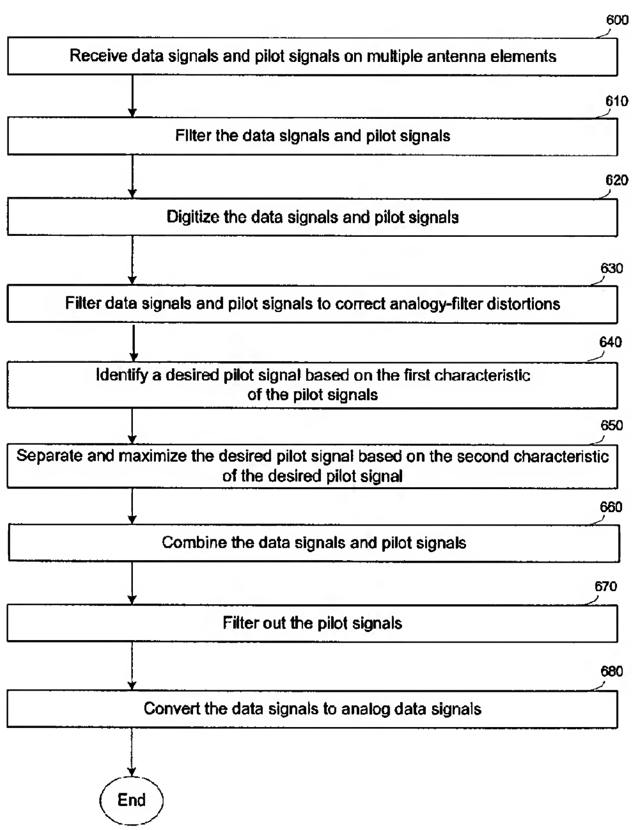
[Continued on next page]

(54) Title: SMART ANTENNA BASED SPECTRUM MULTIPLEXING USING A PILOT SIGNAL



(57) Abstract: A system and method for using a pilot signal in a communication receiver having multiple antenna elements is described. A set of data signals and a set of pilot signals are received (600). A first pilot signal from the set of pilot signals is identified based on a first characteristic of the first pilot signal from the set of pilot signals (640). A set of weight values associated with the antenna elements are adjusted so that a second characteristic of the first pilot signal is substantially optimized with respect to the second characteristic of the remaining pilot signals from the set of pilot signals (650). Consequently, a first data signal from the set of data signals and being uniquely associated with the first pilot signal is substantially optimized by the adjusting of the set of weight values associated with the antenna elements.

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SMART ANTENNA BASED SPECTRUM MULTIPLEXING USING A PILOT SIGNAL

Background

The present invention relates generally to communications and more particularly to a system and method for using a pilot signal added to a transmitted signal in a communication system, and used by the receiving end, in conjunction with multiple antenna elements. The receiver can implement a separation process known as spatial filtering, or also referred to herein as smart antenna.

Broadband networks having multiple information channels are subject to certain types of typical problems such as inter-channel interference and a limited bandwidth per information channel. For example, broadband wireless networks can use cellular and frequency-reuse schemes to extend service areas for a given range of allocated frequencies. In such a broadband wireless network, a large number of different frequency bands are used for the overall system. Adjacent cells are then able to use a different frequency band to minimize interference.

This large number of frequency bands, however, involves an extensive spectrum allocation that can be expensive or difficult. In addition, a limited amount of bandwidth is available for each frequency associated with a given cell.

In sum, a need exists for an improved system and method that can significantly reduce the amount allocated spectrum to communicate a given amount of data or that can significantly increase the amount of data for a given amount of allocated spectrum.

Summary of the Invention

A system and method for using a pilot signal in a communication receiver having multiple antenna elements is described. A set of data signals and a set of pilot signals are received. A first pilot signal from the set of pilot signals is identified based

on a first characteristic of the first pilot signal from the set of pilot signals. A set of weight values associated with the antenna elements are adjusted so that a second characteristic of the first pilot signal is substantially optimized with respect to the second characteristic of the remaining pilot signals from the set of pilot signals. Consequently, a first data signal from the set of data signals and being uniquely associated with the first pilot signal is substantially optimized by the adjusting of the set of weight values associated with the antenna elements.

Brief Description of the Drawings

FIG. 1 shows a system block diagram of a communication system using downlink spectrum multiplexing, according to an embodiment of the invention.

FIG. 2 shows a system block diagram of a communication system using uplink spectrum multiplexing, according to an embodiment of the invention.

FIG. 3 shows a graph of frequency versus amplitude for data signals and pilot signals within an allocated frequency band according to an embodiment of the invention.

FIGS. 4A through 4D show a system block diagram of a transmitter having a pilot transmit subsystem, according to an embodiment of the invention.

FIG. 5 shows a system block diagram of a receiver having a pilot receive subsystem, according to an embodiment of the invention.

FIG. 6 shows a flowchart for receiving and enhancing data signals according to an embodiment of the present invention.

FIG. 7 shows a system block diagram of a pilot-receive subsystem according to an embodiment of the invention.

FIG. 8 shows a flowchart for separating and maximizing the desired pilot signal according to an embodiment of the present invention.

Detailed Description

The disclosed system and method uses a pilot signal to identify and enhance a desired data signal while minimizing undesired data signals. A desired communication source (e.g., a desired basestation) transmits a data signal and a pilot signal. A communication receiver receives the data signal and the pilot signal from the desired communication source and at the same time receives data signals and the pilot signals from undesired communication sources (e.g., undesired basestations). Thus, from the perspective of the communication receiver, it receives data signals and pilot signals where each data signal is uniquely associated with a pilot signal. The communication receiver then identifies the pilot signal from the desired communication source based on a first characteristic of the pilot signal. For example, the first characteristic of the pilot signal can be a unique frequency. The communication receiver, having multiple antenna elements, calculates weight values for each antenna element so that a second characteristic of the desired pilot signal is substantially optimized with respect to the second characteristic of the remaining undesired received pilot signals. The second characteristic of the pilot signals can be, for example, a power level. Accordingly, once the communication receiver has been optimized to receive the desired pilot signal, receiving the desired data signal will also be optimized.

The transmission of the pilot signal can be performed on the uplink and/or the downlink. For example, in a wireless communication system having multiple basestations and multiple handsets, a pilot signal can be transmitted on the downlink from each basestation. In this configuration, a handset receiving signals from multiple basestations can use the pilot signal from the desired basestation to optimize the data signal from that desired basestation. In such a configuration, each handset includes multiple antenna elements. In an alternative configuration, a pilot signal can be transmitted on the uplink from each handset. In this configuration, a basestation receiving signals from multiple handsets can use the pilot signal from the desired handset to optimize the data signal from that desired handset. In this configuration, each basestation includes multiple antenna elements.

Note that embodiments of the invention can be used in wireless or wired communications. For example, an embodiment of the invention can be used in multiple-channel wireless communications using, for example, the WiFi (i.e., the IEEE 802.11A) standard. For another example, an embodiment of the invention can be used in a multiple-channel cable system using, for example, the Data Over Cable Service Interface Specifications (DOCSIS) standard.

FIG. 1 shows a system block diagram of a communication system using downlink spectrum multiplexing, according to an embodiment of the invention. As shown in FIG. 1, network 100 is coupled to basestations 110, 120 and 140, which can in turn be coupled to subscriber unit 130. Note that although FIG. 1 shows three basestations 110, 120 and 140, any number N of basestations can be coupled to network 100. Basestation 110 includes receiver 111 and transmitter 112, which also includes pilot transmit subsystem 113. Basestation 120 includes receiver 121 and transmitter 122, which also includes pilot transmit subsystem 123. Basestation 140 includes receiver 141 and transmitter 142, which also includes pilot transmit subsystem 143. Basestations 110, 120 and 140 can be coupled to subscriber unit 130, for example, by wireless links 150, 152 and 154, respectively. Subscriber unit 130 includes transmitter 132 and receiver 131, which includes pilot receive subsystem 134. In addition, subscriber unit 130 includes a number M of multiple antenna elements that are uncorrelated. In this embodiment, the number N of basestations 110, 120 and 140 can be, for example, greater than the number M of antenna elements at subscriber unit 130.

For the embodiment shown in FIG. 1, downlink spectrum multiplexing is performed by multiple basestations that are transmitting over the same broadband channel frequency band (also referred to herein as a data-frequency band). Each basestation 110, 120 and 140 also transmits a narrowband pilot signal with the broadband modulated data signal. The narrowband pilot signal sent by each basestation 110, 120 and 140 is slightly different from the remaining pilot signals sent by the remaining basestations 110, 120 and 140. In this embodiment, the pilot signals are slightly different from each other in the sense that each pilot signal has an associated frequency band that differs from the frequency bands for the remaining pilot signals.

The subscriber unit 130 uses multiple antenna elements so that the desired broadband signal can be enhanced and the undesired broadband signals can be suppressed. The desired broadband signal originates from the basestation that targets this subscriber. The undesired broadband signals originate from the basestations that do not target this subscriber although they send data signals within the same channel (the same channel defined, for example, by the same time, frequency or code depending on the system configuration). The subscriber's unit 130 suppresses undesired broadband data signals and enhances the desired broadband data signal by monitoring only the different narrowband pilot signals and manipulating the different antenna elements output so that the desired narrowband pilot signals is enhanced while the undesired narrowband pilot signals are suppressed.

In sum, an embodiment using downlink spectrum multiplexing allows multiple basestations each to transmit a narrowband pilot signal with its broadband data signal. The broadband data signal sent by these multiple basestations can be within the same frequency band. Meanwhile, the subscriber units configured to communicate with one or more of these basestations each have multiple antenna elements and a pilot receive subsystem that uses the received pilot signals to enhance the desired data signal.

FIG. 2 shows a system block diagram of a communication system using uplink spectrum multiplexing, according to an embodiment of the invention. As shown in FIG. 2 , network 100 is coupled to basestation 160, which can in turn be coupled to subscriber units 170, 180 and 190. Note that although FIG. 2 shows three subscriber units 170, 180 and 190, any number N of subscriber units can be coupled to basestation 160. Similarly, other basestations (not shown in FIG. 2) can be coupled to network 100. Subscriber unit 170 includes receiver 171 and transmitter 172, which also includes pilot transmit subsystem 173. Subscriber unit 180 includes receiver 181 and transmitter 182, which also includes pilot transmit subsystem 183. Subscriber unit 190 includes receiver 191 and transmitter 192, which also includes pilot transmit subsystem 193. Subscriber units 170, 180 and 190 can be coupled to basestation 160, for example, by wireless links 165, 167 and 169, respectively. Basestation 160 includes transmitter 162 and receiver 161, which includes pilot receive subsystem 164. In addition, basestation 160 includes a number M of multiple antenna elements that are

uncorrelated. In this embodiment, the number N of subscriber units 170, 180 and 190 can be, for example, greater than the number M of antenna elements are basestation 160.

For the embodiment shown in FIG. 2, uplink spectrum multiplexing is performed by multiple subscriber units that are transmitting over the same broadband channel frequency band (also referred to herein as a data-frequency band). Each subscriber units 170, 180 and 190 also transmits a narrowband pilot signal with the broadband modulated data signal. The narrowband pilot signal sent by each subscriber unit 170, 180 and 190 is slightly different from the remaining pilot signals sent by the remaining subscriber units 170, 180 and 190. In this embodiment, the pilot signals are slightly different from each other in the sense that each pilot signal has an associated frequency band that differs from the frequency bands for the remaining pilot signals.

The basestation 160 uses multiple antenna elements so that the desired broadband signal can be enhanced and the undesired broadband signals can be suppressed. The desired broadband signal originates from the subscriber unit that is targeted the basestation 160. The undesired broadband signals originate from the subscriber units that do not target this basestation 160 although they send data signals within the same data-frequency band. The basestation 160 suppresses undesired broadband data signals and enhances the desired broadband data signal by monitoring only the different narrowband pilot signals and manipulating the different antenna elements output so that the desired narrowband pilot signals is enhanced while the undesired narrowband pilot signals are suppressed.

In sum, an embodiment using uplink spectrum multiplexing allows multiple subscriber units each to transmit a narrowband pilot signal with its broadband data signal. The broadband data signal sent by these multiple subscriber units can be within the same frequency band. Meanwhile, the basestation configured to communicate with one or more of these subscriber units has multiple antenna elements and a pilot receive subsystem that uses the received pilot signals to enhance the desired data signal.

FIG. 3 shows a graph of frequency versus amplitude for data signals and pilot signals within an allocated frequency band according to an embodiment of the invention. As shown in FIG. 3, an allocated frequency band 200 includes a data frequency band 210 and pilot-signal bands 220 through 270. The data frequency band 210 uses a portion, for example, 90 percent of the allocated frequency band 200. The remaining portions 280 and 290 of the allocated frequency band 200 are typically used as guard bands (also referred to as being outside of the power-spectrum mask). These remaining portions 280 and 290 can be, for example, a total of 10 percent of the allocation frequency band 200 (i.e., 5 percent on either side of the data frequency band 210). Within these remaining portions 280 and 290 of the allocated frequency band 200, the pilot signals 220 through 270 can be allocated. In sum, the data frequency band 210 can be a broadband channel frequency band and the pilot-signal bands 220 through 270 can be narrowband frequency band.

More specifically, pilot signals 220 through 240 can be allocated within portion 280 and pilot signals 250 through 270 can be allocated within portion 290. For example, pilot signals 220 through 270 each can represent about one percent of the total allocated frequency within allocation frequency band 200. In such a configuration, each pilot signal 220 through 240 can have a signal-to-noise ratio (SNR) similar to the SNR of the data signals within data frequency band 210 without interfering with the data signals within data frequency band 210 or adjacent pilot signals. For example, where the data-frequency band is 90 percent of the allocated-frequency band and each pilot-signal band is one percent of the allocated-frequency band, a corresponding pilot signal has a (10 log(90/1) – 20) dB advantage in SNR.

Thus, following the example of downlink spectrum multiplexing shown in FIG. 1, basestation 110 can send its own data signal within data frequency band 210 and a pilot signal within pilot-signal band 220. Basestation 120 can send its own data signal within data frequency band 210 and a pilot signal within pilot-signal band 230. Basestation 140 can send its own data signal within data frequency band 210 and a pilot signal within pilot-signal band 240. Other basestations not shown in FIG. 1 can send their own data signals within data frequency band 210 and their own pilot signal within pilot-signal bands 250 through 270. Note that the data signals for basestations 110, 120

and 140 and the basestations not shown in FIG. 1 are within and overlap with the data-frequency band 210.

In this configuration, the pilot signals within pilot-signal bands 220 through 270 each have two characteristics that allow for identification and the enhancement of desired data signals. The first characteristic is the frequency of the pilot signal, for example, the center frequency or the specific frequency band. A receiver, such as subscriber unit 130, can know beforehand which basestation is the desired source. The corresponding pilot signal will also then be known. Consequently, a band filter can be used to identify and isolate the desired pilot signal.

The second characteristic of the pilot signals is the power of the pilot signal, for example, the integrated power across the entire pilot-signal band for the desired pilot signal. Again following the example shown in FIG. 1, basestation 110 having a pilot signal within pilot-signal band 220 can be the desired basestation for subscriber unit 130. Consequently, subscriber unit 130 can adjust weight values associated with the antenna elements (not shown in FIG. 1) to maximize the total power of the desired pilot signal within pilot-signal band 220.

Similarly, following the example of uplink spectrum multiplexing shown in FIG. 2, subscriber unit 170 can send its own data signal within data frequency band 210 and a pilot signal within pilot-signal band 220. Subscriber unit 180 can send its own data signal within data frequency band 210 and a pilot signal within pilot-signal band 230. Subscriber unit 190 can send its own data signal within data frequency band 210 and a pilot signal within pilot-signal band 240. Other subscriber units not shown in FIG. 2 can send their own data signals within data frequency band 210 and their own pilot signal within pilot-signal bands 250 through 270. Note that the data signals for subscriber units 170, 180 and 190 and the subscriber units not shown in FIG. 2 are within and overlap with the data-frequency band 210.

In this configuration, the pilot signals within pilot-signal bands 220 through 270 each have two characteristics that allow for identification and the enhancement of desired data signals. The first characteristic is the frequency of the pilot signal, for

example, the center frequency or the specific frequency band. A receiver, such as basestation 160, can know beforehand which subscriber unit is the desired source. The corresponding pilot signal will also then be known. Consequently, a band filter can be used to identify and isolate the desired pilot signal.

The second characteristic of the pilot signals is the power of the pilot signal, for example, the integrated power across the entire pilot-signal band for the desired pilot signal. Again following the example shown in FIG. 2, subscriber unit 170 having a pilot signal within pilot-signal band 220 can be the desired subscriber unit for basestation 160. Consequently, basestation 160 can adjust weight values associated with the antenna elements (not shown in FIG. 2) to maximize the total power of the desired pilot signal within pilot-signal band 220.

Note that the example shown in FIG. 3 is merely for explanatory purposes. Any other configurations, the specific size of the allocated frequency band 200, the data frequency band 210 and the pilot-signal bands 220 through 270 can be different. In addition, the example shown in FIG. 3 is based on a specific embodiment where the pilot signals are narrow band signals within the guard band. Many other types of embodiments are possible where the pilot signals, for example, are within the data frequency band of a spread spectrum system, created as an artificial multipath, embedded within an orthogonal frequency division multiplexing (OFDM) system, etc. These and other examples of different embodiments are discussed below after the discussion of Figures 4 through 8 in connection with the narrow-band pilot-signal example.

FIGS. 4A through 4D shows a system block diagram of a transmitter having a pilot transmit subsystem, according to an embodiment of the invention. By way of illustration, FIGS. 4A through 4D show a system block diagram of transmitters 300, 310, 320 and 330. Any of these transmitters 300, 310, 320 and 330 can correspond to the any of the tranmitters 112, 122 and 142 of FIG. 1 and transmitters 172, 182 and 192 of FIG. 2.

As shown in FIG. 4A, transmitter 300 includes transmitter baseband module 301, pilot transmit subsystem 308, modulator 302, intermediate frequency (IF) module 303, radio frequency (RF) module 304 and antenna elements 305. These components are coupled in series. Pilot transmit subsystem 308 includes digital adder 306, which receives a digital pilot signal 307. The data signal to be transmitted by transmitter 300 is provided from transmitter baseband module 301 to digital adder 306. This data signal is in digital form. The digital adder 306 adds digital pilot signal 307 to the digital data signal. The digital data signal and digital pilot signal are converted to analog signals by modulator 302. The frequencies of these analog signals are converted from baseband frequencies to intermediate frequencies by IF module 303. The frequencies of these signals are then converted to radio frequencies by RF module 304. These signals are then transmitted by antenna elements 305.

As shown in FIG. 4B, transmitter 310 includes transmitter baseband module 311, modulator 312, pilot transmit subsystem 318, IF module 313, RF module 314 and antenna elements 315. These components are coupled in series. Pilot transmit subsystem 318 includes adder 316, which receives an analog pilot signal 317. The data signal to be transmitted by transmitter 310 is provided from transmitter baseband module 311 to modulator 312. The digital data signal is converted to an analog signal by modulator 312. The digital signal is provided to adder 316, which adds the analog pilot signal 317. The frequencies of these analog signals are converted from baseband frequencies to intermediate frequencies by IF module 313. The frequencies of these signals are then converted to radio frequencies by RF module 314. These signals are then transmitted by antenna elements 315.

As shown in FIG. 4C, transmitter 320 includes transmitter baseband module 321, modulator 322, IF module 323, pilot transmit subsystem 328, RF module 324 and antenna elements 325. These components are coupled in series. Pilot transmit subsystem 328 includes adder 326, which receives an analog pilot signal 327. The data signal to be transmitted by transmitter 320 is provided from transmitter baseband module 321 to modulator 322. The digital data signal is converted to an analog signal by modulator 322. The frequencies of this analog data signal are converted from baseband frequencies to intermediate frequencies by IF module 323. The analog data

signal is provided to adder 326, which adds the analog pilot signal 327. The frequencies of these signals are then converted to radio frequencies by RF module 324. These signals are then transmitted by antenna elements 325.

As shown in FIG. 4D, transmitter 330 includes transmitter baseband module 331, modulator 332, IF module 333, RF module 334, pilot transmit subsystem 338 and antenna elements 335. These components are coupled in series. Pilot transmit subsystem 338 includes adder 336, which receives an analog pilot signal 337. The data signal to be transmitted by transmitter 330 is provided from transmitter baseband module 331 to modulator 332. The digital data signal is converted to an analog signal by modulator 332. The frequencies of this analog data signal are converted from baseband frequencies to intermediate frequencies by IF module 333. The frequencies of this analog data signal are then converted to radio frequencies by RF module 334. The analog data signal is provided to adder 336, which adds the analog pilot signal 337. These signals are then transmitted by antenna elements 335.

FIG. 5 shows a system block diagram of a receiver having a pilot receive subsystem, according to an embodiment of the invention. The embodiment shown in FIG. 5 can correspond to the receiver 131 of FIG. 1 and receiver 161 of FIG. 2. Note that although FIG. 5 shows a specific embodiment of a receiver having four antenna elements, a receiver can have any number of two or more antenna elements. Such a receiver will have component sets that correspond to the specific number of antenna elements for that receiver embodiment.

As shown in FIG. 5, receiver 500 includes antenna elements 501, 502, 503 and 504, which are coupled to filters 511, 512, 513 and 514, respectively. Filters 511, 512, 513 and 514 are coupled to A/D converters 521, 522, 523 and 524, respectively, which in turn are coupled to software filters 531, 532, 533 and 534, respectively. Software filters 531 through 534 are coupled to pilot receive subsystem 540, which is also coupled to digital signal processor 550 and combiner 560. Combiner 560 is coupled to filter 570, which in turn is coupled to D/A converter 580. For illustrative purposes, the operation of receiver 500 will be explained in reference to the flow chart of FIG. 6.

FIG. 6 shows a flowchart for receiving and enhancing data signals according to an embodiment of the present invention. At step 600, data signals and pilot signals are received on multiple antenna elements. The data signals and pilot signals can be received separately, for example, on antenna elements 501 through 504 as shown in FIG. 5. Thus, each antenna element will generate a composite of the data signals and pilot signals received at its given location.

At step 610, each filter (i.e., 511 to 514) filters the data signals and pilot signals received by its associated antenna elements. As shown in FIG. 5, these data signals and pilot signals can be filtered at filters 511 through 514. These filters can be, for example, hardware filters that filter the signals while in an analog form. At step 620, the filtered data signals and pilot signals are digitized. The data signals and pilot signals can be digitized, for example, by A/D converters 521 through 524. In other words, the signals from filters 511 through 514 are provided to A/D converters 521 through 524, respectively, which digitize each set of signals.

At 630, the digitized data signals and pilot signals are filtered in software to correct for distortions due to the hardware filters 511 through 514. In other words, software filters 531, 532, 533 and 534 correct for distortions that were induced by filters 511, 512, 513 and 514, respectively.

At step 640, the desired pilot signal is identified based on the first characteristic of the pilot signals. As shown in FIG. 5, digital signal processor 550 can identify the desired pilot signal from the pilot signals stored in pilot-receive subsystem 540 by, for example, a specific frequency or frequency band of the desired pilot signal. Note that digital signal processor 550 can also provide the appropriate control / status signals to the components of the pilot-receive subsystem 540 via connections not shown in FIG. 5. These components of pilot-receive subsystem 540 are described in further detail below in reference to FIG. 7.

At step 650, the desired pilot signal is separated and maximized based on the second characteristic of the desired pilot signal. For example, the second characteristic of the desired pilot signal can be the total power within the power-signal

band for that desired pilot signal. Based on this second characteristic, the pilot-receive subsystem 540 can then adjust the weight values associated with antenna elements 501 through 504 so that the power across the pilot-signal band for the desired pilot signal is maximized. Consequently, the power across the pilot-signal bands for the remaining pilot signals (i.e., the undesired pilot signals) will be minimized by this process. Also, because enhancing the desired pilot signal corresponds to changes in the desired data signal, the desired data signal will also be maximized in the process of maximizing the desired pilot signal.

At step 660, the data signals and pilot signals output from the pilot-receive subsystem 540 are combined by combiner 560. More specifically, the pilot-receive subsystem 540 produces a set of outputs (each having data signals and pilot signals) where each output uniquely corresponds to an antenna element 501 through 504. Combiner 560 combines this set of outputs into a single output having the data signals and pilot signals corresponding to all of the antenna elements 501 through 504.

At step 670, the pilot signals are filtered out so that only the data signals remain. Turning to FIG. 5, the pilot signals can be filtered out by filter 570, which can be for example a band-pass filter. Following the example shown in FIG. 3, the band-pass filter 570 can correspond to the data frequency band 210 shown in FIG. 3. Correspondingly, the pilot signals within pilot signal bands 220 through 270 are removed by band-pass filter 570.

At step 680, the data signals (in digital form) are converted to analog signals 590. Note that the analog data signals 590 produced by D/A converter 580 represents the desired data signal as well as the undesired data signals. Due to the maximization process performed by pilot-receive subsystem 540, however, the desired data signals are maximized or enhanced while the remaining data signals (i.e., undesired data signals) are minimized. Consequently, these undesired data signals interfere with the desired data signal less and the desired data signal is enhanced.

FIG. 7 shows a system block diagram of a pilot-receive subsystem according to an embodiment of the invention. More specifically, the pilot-receive

subsystem 700 shown in FIG. 7 corresponds to the pilot-receive subsystem 540 shown in FIG. 5.

Pilot-receive subsystem 700 includes circuits 710. Note that although only one circuit 710 is shown in FIG. 7, multiple circuits are present within pilot-receive subsystem 700. The specific number of circuits 710 corresponds to the specific number of antenna elements (e.g., antenna elements 501 through 504 shown in FIG. 5). Thus, for the receiver shown in FIG. 5 and having four antenna elements, pilot-receive subsystem 700 consequently has four circuits 710.

Circuit 710 includes four-port memory 711, which is coupled to filter 712, filter 713, and complex weight module 714. Filter 712 is coupled to filter 715, which is in turn coupled to memory storages 716 and 717. Memory storages 716 and 717 are coupled to weight-application modules 718 and 719, respectively.

The weight-application modules 718 and 719 for each circuit 710 are coupled to best solution selector 720, which is in turn coupled to weight calculator 730. Weight calculator 730 is also coupled to weight-application modules 718 and 719 from each of the circuits 710. Best solution selector 720 also outputs a value 725 when a best solution for the weight values is obtained. This value 725 is also provided to the complex-weight module 714 of every circuit 710. The operation of pilot-receive subsystem 700 will now be described with reference to the flowchart shown in FIG. 8.

FIG. 8 shows a flowchart for separating and maximizing the desired pilot signal according to an embodiment of the present invention. More specifically, the flowchart shown in FIG. 8 corresponds to step 650 shown in FIG. 6 and typically is performed by pilot-receive subsystem 540 shown in FIG. 5.

Note that steps 800 through 845 shown in FIG. 8 are performed in parallel by separate circuits 710 each of which uniquely corresponds to an antenna element 501 through 504 shown in FIG. 5. More generally, the number of circuits will correspond to the number of antenna elements of a given receiver. Thus, although steps 800 through 845 are discussed in reference to a single circuit 710, the same steps are performed in parallel for all of the circuits 710.

At step 800, the data signals and pilot signals are stored. These data signals and pilot signals can be in digital form and received from one of the software filters 531 through 534 shown in FIG. 5. At step 805, the stored data signals and pilot signals are filtered to produce a reduced number of pilot-signal samples. In other words, the data signals within data frequency band 210 as shown in FIG. 3 are removed and only the pilot signals within pilot-signal bands 220 through 270 in FIG. 3 remain.

At step 810, these pilot signals are further filtered to produce an in-phase component and a quadature component. At step 820, the in-phase component is stored. At step 825, the quadrature component is stored. Turning to FIG. 7, the in-phase component produced by filter 715 is stored in memory storage 716, and the quadrature component produced by filter 715 is stored in memory storage 717. The stored pilot-signal samples stored in memory storage 716 and 717 can be used iteratively to determine the appropriate weight values associated with the in-phase and quadature signals.

The specific number of pilot-signal samples is related to the bandwidth of pilot-signal band. More specifically, the pilot-signal samples for a given pilot signal are collected over a time period on the order of 1/B, where B is the bandwidth of the pilot-signal band for that pilot signal. Consequently, the number of samples needed is relatively small. For example, for a pilot-signal band having a bandwidth of 20 kHz, the number of samples should be on the order of one per 25 μsec. Thus, for a message that is only 100 μsec long, only 4 samples are needed.

At step 830, the power for each pilot-signal band corresponding to the in-phase component is calculated. At step 835, the power for each pilot-signal band for the quadrature component is calculated.

At step 840, a weight value for the in-phase component is calculated. At step 845, a weight value for the quadrature component is calculated. Steps 830 and 840 can be performed by weight-application module 718. Steps 835 and 845 can be performed by weight-application module 719. The weight values can be calculated, for

example, by applying a gradient descent to the power for each pilot-signal band calculated in steps 830 and 835.

At step 850, the power for each pilot-signal band is calculated for all antenna elements. More specifically, best solution selector 720 receives the weight values for the in-phase and quadature components from each circuit 710 and then determines the power for each pilot-signal band based on these new weight values. In other words, the best solution selector 720 receives two weight values (one for the in-phase component and the other for the quadrature component) for each circuit 710. Then, using these weight values (two weight values times the number of circuits 710), determines power for each pilot-signal band using these new weight values.

At conditional step 855, a determination is made as to whether to continue with another iteration of calculating weight values. For example, the iterations can continue until a maximum difference between the desired pilot-signal bands and the remaining pilot-signal bands has been achieved. Alternatively, the iterations can continue until a maximum number of iterations have been performed. Performing additional iterations allow the receiver possibly to obtain a new set of weight values that better enhance the desired pilot signal while suppressing the undesired pilot signals. If the iterations are to continue, then the process proceeds to step 860. At step 860, weight values for the next iteration are selected. Weight calculator 730 can perform step 860. These newly selected weight values are then provided to the weight-application modules 718 and 719 for every circuit 710.

Returning to FIG. 8, after weight values for the next iteration are selected at steps 860, the process continues at steps 830 and 835 where the power for each pilot-signal band is calculated based on these newly selected weight values. The process then continues for steps 830 and 835 through to 855 until a determination is made that the iterations should no longer continue. At this point, the process proceeds to step 870. Note that the iterations can be performed at a relatively slow rate of, for example, one iteration per sample (i.e., comparing one sample to another during an iteration). Thus, for a configuration using six pilot-signal bands (and having six circuits 710), for

example, two pilot-signal samples for each pilot-signal band, totaling 12 pilot-signal samples, can be compared for a given iteration by each circuit 710.

Steps 870 and 880 are performed for each antenna element. In other words, steps 870 and 880 are performed in parallel by each circuit 710. At step 870, the weight-adjusted data signals and pilot signals (using the final weight values 725) are filtered to find the beginning and end times of the data signals. Returning to FIG. 7, the final weight values 725 are provided to filter 713. Filter 713 also receives the original data signals and filter signals. Filter 713 uses the weight-adjusted values of the data signals and pilot signals to determine the precise beginning and end of the message within the desired data signal. These beginning and end times are then provided to complex-weight module 714.

At step 880, the final weight values are applied to the original data signals and pilot signals to produce an enhanced desired data signal. Returning to FIG. 7, complex-weight module 714 receives from four-port memory 711 a copy of the original data signals and pilot signals and also receives the final weight values 725 from best solution selector 720. Using the beginning and end times provided by filter 713, complex-weight module 714 then applies the final weight values to the original data signals and pilot signals and produces output data 740. As discussed above, now that the receiver has been optimized to receive the desired pilot signal, the desired data signal is also now optimized.

Although the above discussion of FIGS. 2 through 8 is based on specific embodiments where the pilot signals are narrow band signals within the guard band, many other types of embodiments are possible. For example, where the number of pilot signals within the system exceeds the number of narrow bands available within the guard bands, the pilot signals can be modulated with a code. Thus, two or more pilot signals within a given pilot-signal band can be identified by the modulation code. In this embodiment, the first characteristic of the pilot signals is a combination of the frequency and the modulation code.

In another embodiment, pilot signals can be included within the data-frequency band for a code-division multiple access (CDMA) system (i.e., a spread spectrum system). In such an embodiment, the pilot signals each can use a spread-spectrum pseudo-noise sequence for identification. The power in the spread-spectrum signal band (i.e., the data-frequency band based on the desired spread-spectrum pseudo-noise sequence) can be used to optimize the desired pilot signal. Thus, for this embodiment, the first pilot-signal characteristic is a spread-spectrum pseudo-noise sequence and the second pilot-signal characteristic is power in the spread-spectrum signal band.

In another embodiment, the pilot signals can be created as an artificial multipath. More specifically, the pilot signals can be included within the data-frequency band and can be generated with a time delay unique for that communication source. The specific time delay can be used as a pilot-signal identifier. The power of the delayed signal can be used to optimize the desired pilot signal. Thus, for this embodiment, the first pilot-signal characteristic is the amount of time delay for the pilot signal and the second pilot-signal characteristic is power of the desired pilot signal.

In yet another embodiment, the pilot signals can be embedded within an orthogonal frequency division multiplexing (OFDM) system. In this embodiment, the unused OFDM carriers can be modulated as pilot signals. Thus, for this embodiment, the first pilot-signal characteristic is the frequency of the pilot signals (e.g., the OFDM center frequencies for the pilot-signal bands) and the second pilot-signal characteristic is power of the desired pilot signal.

In yet another embodiment, the pilot signals can be within the data-frequency band and amplitude modulated with a unique code. Thus, for this embodiment, the first pilot-signal characteristic is the amplitude-modulation code of the pilot signals and the second pilot-signal characteristic is power of the desired amplitude-modulated pilot signal.

In yet another embodiment, the pilot signals can be within the data-frequency band and frequency shifted with a unique code. Thus, for this embodiment,

the first pilot-signal characteristic is the frequency-shifted code of the pilot signals and the second pilot-signal characteristic is power of the desired frequency-shifted pilot signal.

In yet another embodiment, the pilot signals can be within the data-frequency band and phase shifted with a unique code. Thus, for this embodiment, the first pilot-signal characteristic is the phase-shifted code of the pilot signals and the second pilot-signal characteristic is power of the desired phase-shifted pilot signal.

Conclusion

While various embodiments of the invention have been described above, it should be understood that they have been presented by way of example only, and not limitation. Thus, the breadth and scope of the invention should not be limited by any of the above-described embodiments, but should be defined only in accordance with the following claims and their equivalents.

The previous description of the embodiments is provided to enable any person skilled in the art to make or use the invention. While the invention has been particularly shown and described with reference to embodiments thereof, it will be understood by those skilled in the art that various changes in form and details may be made therein without departing from the spirit and scope of the invention.

For example, although FIG. 3 above describes an embodiment for narrow-band pilot signals for particular system parameters. Other types of system parameters are possible. For example, an embodiment can be configured within a multiple-channel cable system using the Data Over Cable Service Interface Specifications (DOCSIS) standard. For such an embodiment, the downlink can use a modulation of 64 QAM. Thus, within an assigned band of 6 MHz, the data-frequency band can use 5.4 MHz. A pilot-signal band can be 100KHz and placed 150 kHz from the side of the data-frequency band. The pilot signal can have the substantially same power as the data signals without degrading the quality of the data signals.

For another example, an embodiment can be configured within a multiple-channel wireless communication system using, for example, the WiFi (i.e., the IEEE 802.11A) standard. Under this standard, the data-frequency band is divided into 64 quality-width channels but only 52 of these channels are actually used for data signals. Consequently, about 15 percent of the data-frequency band is unused by the data signals. Accordingly, pilot signals can be located on these unused channels.

For another example, an embodiment can be configured within a communication system according to the Broadband Wireless Internet Forum (BWIF) standard. Under this standard, consider an example of 128 channels within the data-frequency band. For this example, only 106 channels are used for data signals while the remaining 22 channels are zero-tone channels. Consequently, about 17 percent of the data-frequency band is unused by the data signals and pilot signals can be located on these unused channels.

What is claimed is:

1. A method for using a pilot signal to enhance a data signal associated with the pilot signal, comprising:

receiving a plurality of data signals and a plurality of pilot signals on a plurality of antenna elements, each data signal from the plurality of data signals being uniquely associated with a pilot signal from the plurality of pilot signals, each pilot signal from the plurality of pilot signals having a first characteristic and a second characteristic;

identifying a first pilot signal from the plurality of pilot signals based on the first characteristic of the first pilot signal; and

adjusting a first weight value associated with each antenna element from the plurality of antenna elements so that the second characteristic of the first pilot signal is substantially optimized with respect to the second characteristic of the remaining pilot signals from the plurality of pilot signals.

2. The method of claim 1, further comprising:

modifying the data signal associated with the first pilot based on the first weight value and a second weight value associated with each antenna element from the plurality of antenna elements to produce a modified data signal;

modifying a transmission data signal based on the first weight value and the second weight value associated with each antenna element from the plurality of antenna elements; and

transmitting the modified transmission data signal.

3. The method of claim 1, further comprising:

performing the following for each antenna element from the plurality of antenna elements:

storing a plurality of signal samples for the first pilot signal; and

filtering the plurality of signal samples for the first pilot signal to produce a plurality of in-phase signal samples and a plurality of quadature signal samples, the first weight value being associated with the plurality of in-phase signal samples, a second weight value being associated with the plurality of quadature signal samples; and

iteratively adjusting the first weight value and the second weight value associated with each antenna element from the plurality of antenna elements so that the second characteristic of the first pilot signal is substantially optimized with respect to the second characteristic of the remaining pilot signals from the plurality of pilot signals.

4. The method of claim 3, further comprising:

scanning, for each antenna element from the plurality of antenna elements, the stored plurality of signal samples for the first pilot signal to produce an indication of a beginning and an end of the data signal associated with the first pilot signal; and

initially applying the first weight value to the data signal associated with the first pilot signal at the beginning indication.

5. The method of claim 1, wherein:

the plurality of data signals is associated with a data frequency band within an allocated frequency band;

the plurality of pilot signals each is uniquely associated with a pilot-signal band within the allocated frequency band and outside the data frequency band;

the first characteristic of each pilot signal from the plurality of pilot signals is at least one from the group of: (a) a frequency of an unmodulated carrier wave and (b) a modulation and a frequency of a modulated carrier wave; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

6. The method of claim 1, wherein:

the plurality of data signals is associated with a data frequency band;

the plurality of pilot signals is associated with the data frequency band;

the first characteristic of each pilot signal from the plurality of pilot signals is a spread spectrum pseudo noise sequence; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power in spread spectrum associated with that pilot signal.

7. The method of claim 1, wherein:

the plurality of data signals is associated with a data frequency band;

the plurality of pilot signals is associated with the data frequency band, each pilot signal being associated with its own time delay from the associated data signal from the plurality of data signals;

the first characteristic of each pilot signal from the plurality of pilot signals is the associated time delay; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

8. The method of claim 1, wherein:

each data signal from the plurality of data signals is uniquely associated with a frequency from a plurality of frequencies;

each pilot signal from the plurality of pilot signals is uniquely associated with a modulation code, each pilot signal from the plurality of pilot signals is uniquely associated with the a remaining frequency from the plurality of frequencies;

the first characteristic of each pilot signal from the plurality of pilot signals is the modulated code; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

9. The method of claim 1, wherein:

each data signal from the plurality of data signals is amplitude modulated with a unique pilot signal having an associated amplitude-modulation code and a power;

the first characteristic of each pilot signal from the plurality of pilot signals is the amplitude-modulation code associated with that pilot signal; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

10. The method of claim 1, wherein:

each data signal from the plurality of data signals is frequency-shift modulated with a unique pilot signal having an associated frequency-shift code and a power;

the first characteristic of each pilot signal from the plurality of pilot signals is the frequency-shift code associated with that pilot signal; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

11. The method of claim 1, wherein:

each data signal from the plurality of data signals is phase-shift modulated with a unique pilot signal having an associated phase-shift code and a power;

the first characteristic of each pilot signal from the plurality of pilot signals is the phase-shift code associated with that pilot signal; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

12. An apparatus having a plurality of antenna elements configured to receive a plurality of data signals and a plurality of pilot signals, each data signal from the plurality of data signals being uniquely associated with a pilot signal from the plurality of pilot signals, each pilot signal from the plurality of pilot signals having a first characteristic and a second characteristic, comprising:

a plurality of circuits each coupled to an antenna element from the plurality of antenna elements, each circuit having:

a filter, the filter configured to receive the plurality of data signals and the plurality of pilot signals, the filter configured to produce a first signal component and a second component;

a first weight-application module coupled to the filter, the first weight-application module configured to receive the first signal component and to apply a first weight value to the first signal component; and

a second weight-application module coupled to the filter, the second weight-application module configured to receive the second signal component and to apply a second weight value to the second signal component;

a processor coupled to the plurality of circuits, the processor configured to determine a first pilot signal from the plurality of pilot signals based on the first characteristic of the first pilot signal; and

a best solution selector coupled to the first weight-application module and the second weight-application module of each circuit from the plurality of circuits, the best

solution selector configured to select an iteration value for the first weight value and the second weight value based on the second characteristic of the pilot signal.

13. The apparatus of claim 12, wherein:

the first weight-application module further configured to calculate the first weight value based on the first signal component so that the second characteristic of the first pilot signal is substantially optimized with respect to the second characteristic of the remaining pilot signals from the plurality of pilot signals;

the second weight-application module further configured to calculate the second weight value based on the second signal component so that the second characteristic of the first pilot signal is substantially optimized with respect to the second characteristic of the remaining pilot signals from the plurality of pilot signals.

14. The apparatus of claim 12, where each circuit from the plurality of circuits further includes:

a second filter coupled to the best solution selector, the filter configured to receive a final value for the first weight value and a final value the second weight value from the best solution selector, the second filter configured to identify a start indicator and an end indicator of the data signal from the plurality of data signals associated with the first pilot signal; and

a complex-weight module coupled to the second filter, the complex-weight module configured to receive the start indicator, the end indicator, the final value of the first weight value and the final value of the second weight value.

15. The apparatus of claim 12, wherein:

the plurality of data signals is associated with a data frequency band within an allocated frequency band;

the plurality of pilot signals each is uniquely associated with a pilot-signal band within the allocated frequency band and outside the data frequency band;

the first characteristic of each pilot signal from the plurality of pilot signals is at least one from the group of: (a) a frequency of an unmodulated carrier wave and (b) a modulation and a frequency of a modulated carrier wave; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

16. The apparatus of claim 12, wherein:

the plurality of data signals is associated with a data frequency band;

the plurality of pilot signals is associated with the data frequency band;

the first characteristic of each pilot signal from the plurality of pilot signals is a spread spectrum pseudo noise sequence; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power in spread spectrum associated with that pilot signal.

17. The apparatus of claim 12, wherein:

the plurality of data signals is associated with a data frequency band;

the plurality of pilot signals is associated with the data frequency band, each pilot signal being associated with its own time delay from the associated data signal from the plurality of data signals;

the first characteristic of each pilot signal from the plurality of pilot signals is the associated time delay; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

18. The apparatus of claim 12, wherein:

each data signal from the plurality of data signals is uniquely associated with a frequency from a plurality of frequencies;

each pilot signal from the plurality of pilot signals is uniquely associated with a modulation code, each pilot signal from the plurality of pilot signals is uniquely associated with the a remaining frequency from the plurality of frequencies;

the first characteristic of each pilot signal from the plurality of pilot signals is the modulated code; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

19. The apparatus of claim 12, wherein:

each data signal from the plurality of data signals is amplitude modulated with a unique pilot signal having an associated amplitude-modulation code and a power;

the first characteristic of each pilot signal from the plurality of pilot signals is the amplitude-modulation code associated with that pilot signal; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

20. The apparatus of claim 12, wherein:

each data signal from the plurality of data signals is frequency-shift modulated with a unique pilot signal having an associated frequency-shift code and a power;

the first characteristic of each pilot signal from the plurality of pilot signals is the frequency-shift code associated with that pilot signal; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

21. The apparatus of claim 12, wherein:

each data signal from the plurality of data signals is phase-shift modulated with a unique pilot signal having an associated phase-shift code and a power;

the first characteristic of each pilot signal from the plurality of pilot signals is the phase-shift code associated with that pilot signal; and

the second characteristic of each pilot signal from the plurality of pilot signals is a power associated with that pilot signal.

22. A method for using a pilot signal in a communication receiver having a plurality of antenna elements, comprising:

receiving a plurality of data signals and a plurality of pilot signals;

identifying a first pilot signal from the plurality of pilot signals based on a first characteristic of the first pilot signal from the plurality of pilot signals; and

adjusting a plurality of weight values associated with the plurality of antenna elements so that a second characteristic of the first pilot signal is substantially optimized with respect to the second characteristic of the remaining pilot signals from the plurality of pilot signals,

whereby a first data signal from the plurality of data signals and being uniquely associated with the first pilot signal is substantially optimized by the adjusting of the plurality of values associated with the plurality of antenna elements.

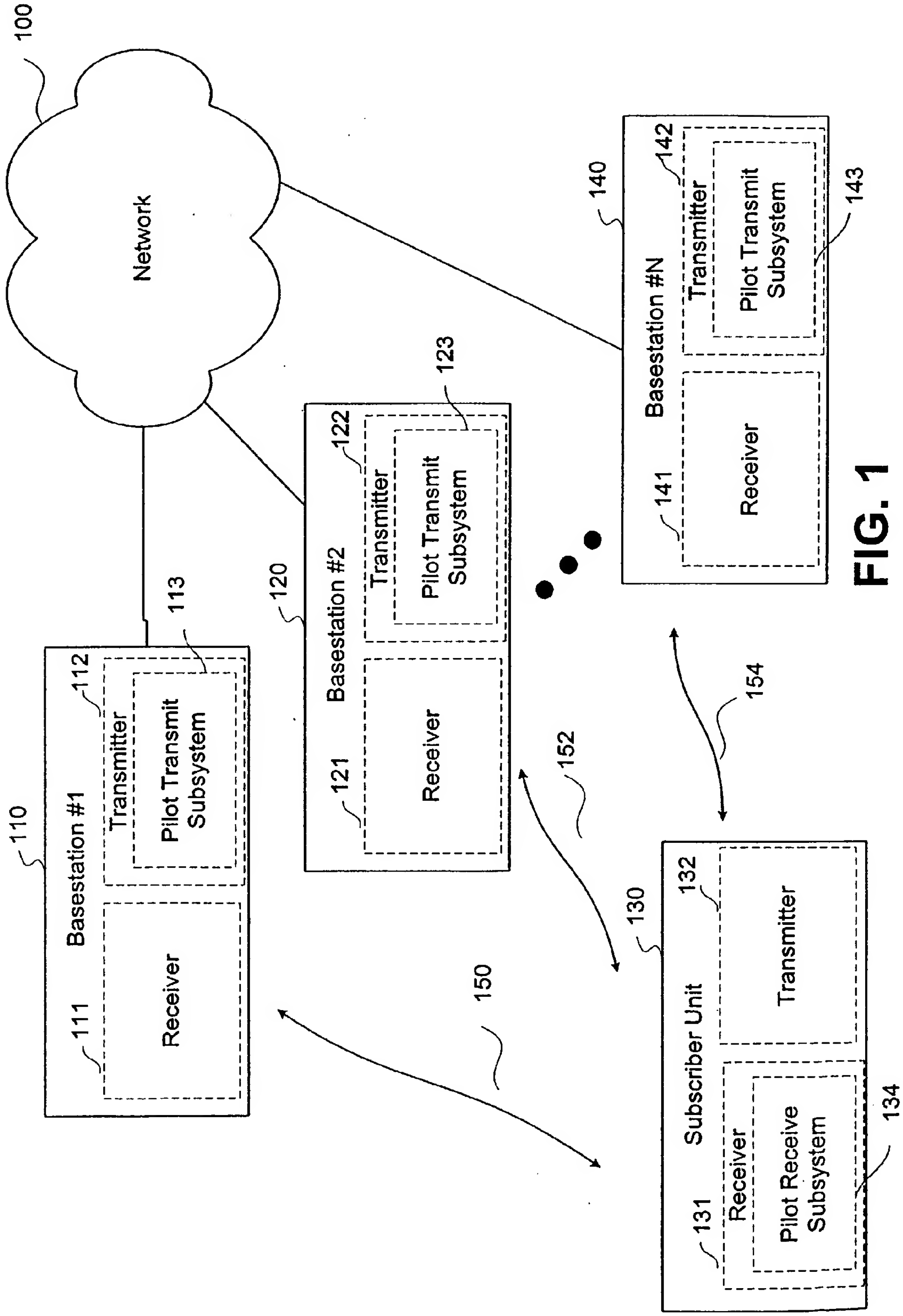
100
Network
110
Basestation #1
111
Receiver
112
Transmitter
113
Pilot Transmit Subsystem
120
Basestation #2
121
Receiver
122
Transmitter
123
Pilot Transmit Subsystem
140
Basestation #N
141
Receiver
142
Transmitter
143
Pilot Transmit Subsystem
130
Subscriber Unit
131
Receiver
134
Pilot Receive Subsystem
132
Transmitter
150
152
154

FIG. 1

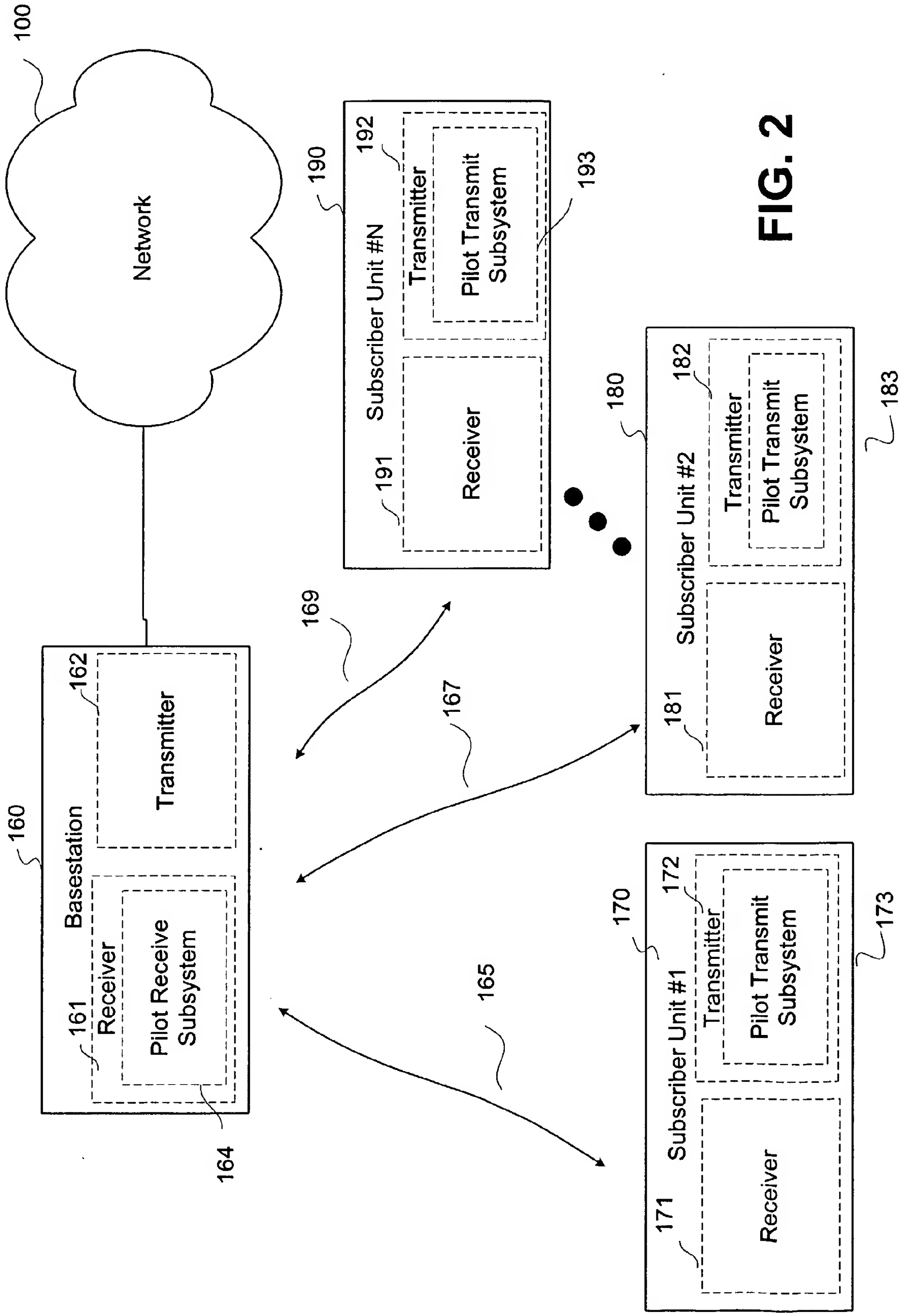
100
Network
160
Basestation
161
Receiver
Pilot Receive Subsystem
164
162
Transmitter
165
167
169
170
Subscriber Unit #1
171
Receiver
172
Transmitter
Pilot Transmit Subsystem
173
180
Subscriber Unit #2
181
Receiver
182
Transmitter
Pilot Transmit Subsystem
183
190
Subscriber Unit #N
191
Receiver
192
Transmitter
Pilot Transmit Subsystem
193

FIG. 2

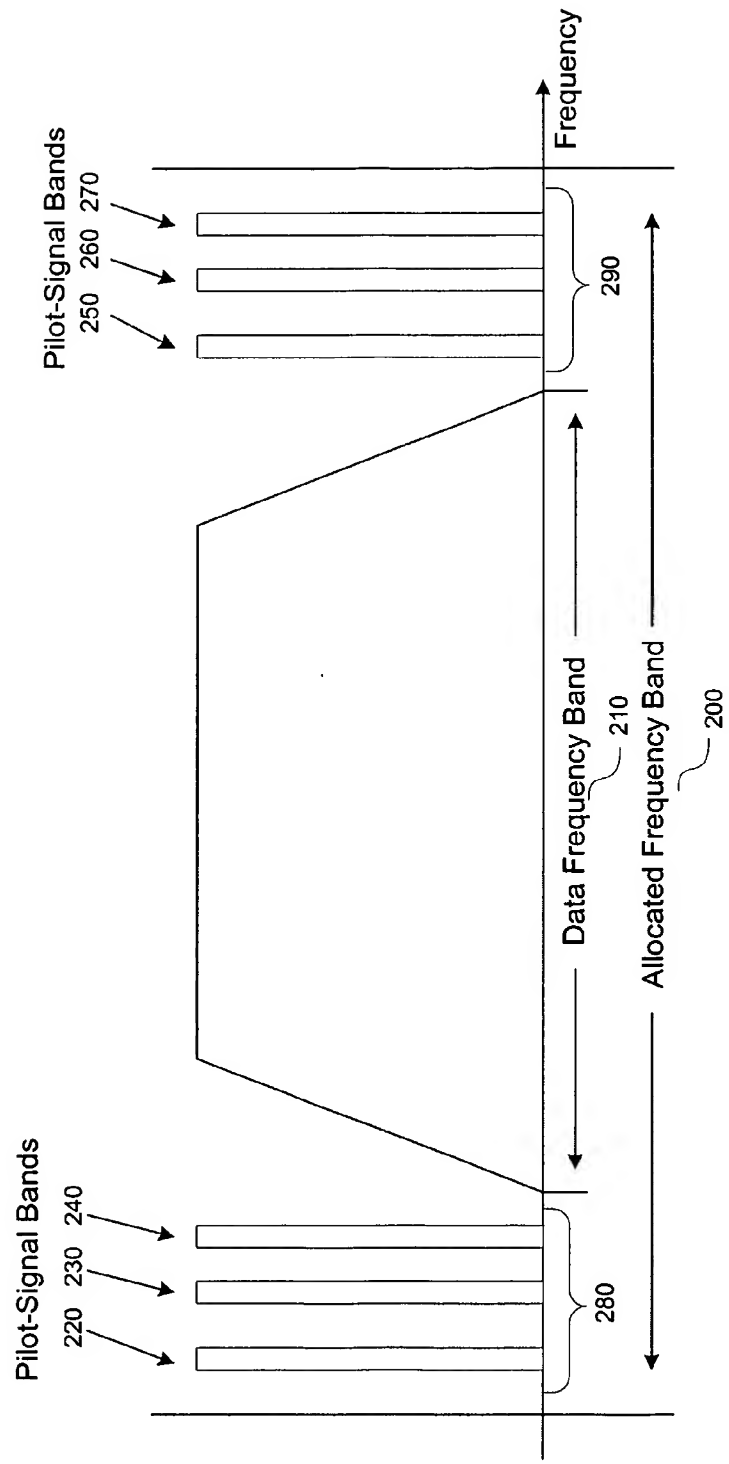
Frequency
Pilot-Signal Bands
250
260
270
290
Data Frequency Band
210
Allocated Frequency Band
200
Pilot-Signal Bands
220
230
240
280

FIG. 3

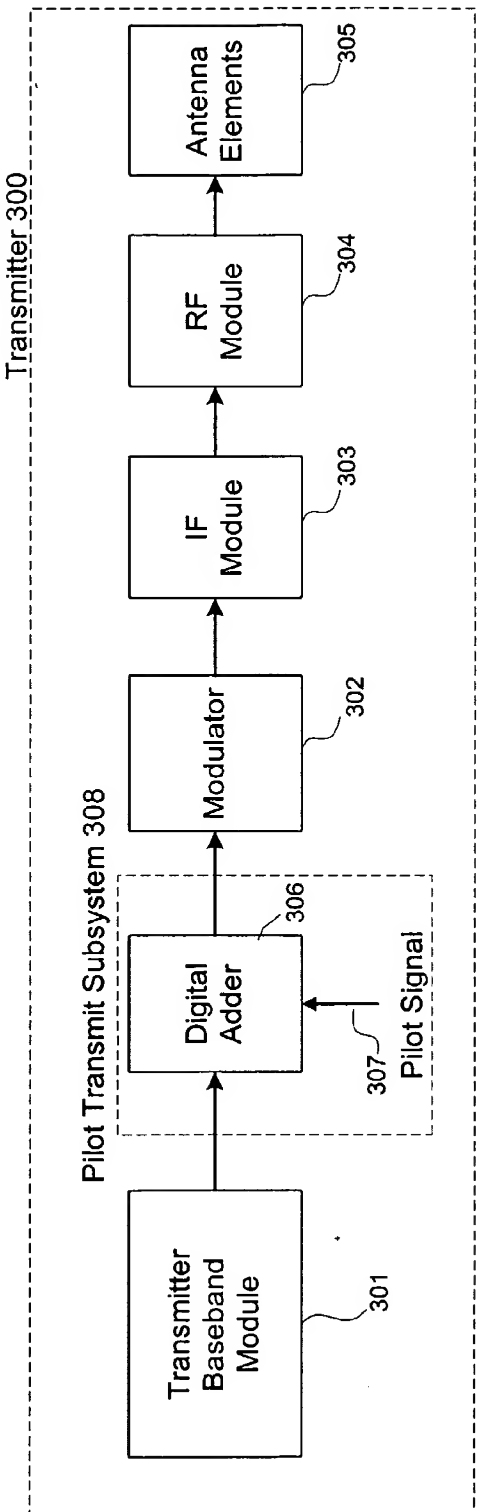
Transmitter 300
Pilot Transmit Subsystem 308
Transmitter Baseband Module
301
Digital Adder
306
307
Pilot Signal
Modulator
302
IF Module
303
RF Module
304
Antenna Elements
305

FIG. 4A

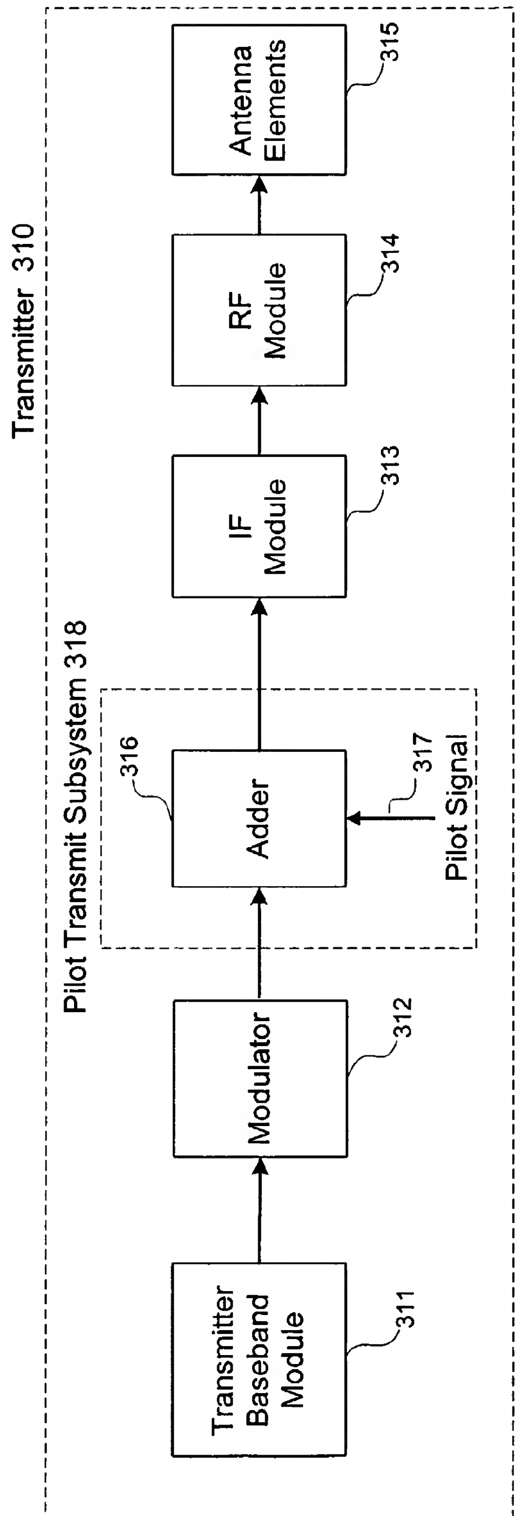
Transmitter 310
Pilot Transmit Subsystem 318
Transmitter Baseband Module
311
Modulator
312
Adder
316
317
Pilot Signal
IF Module
313
RF Module
314
Antenna Elements
315

FIG. 4B

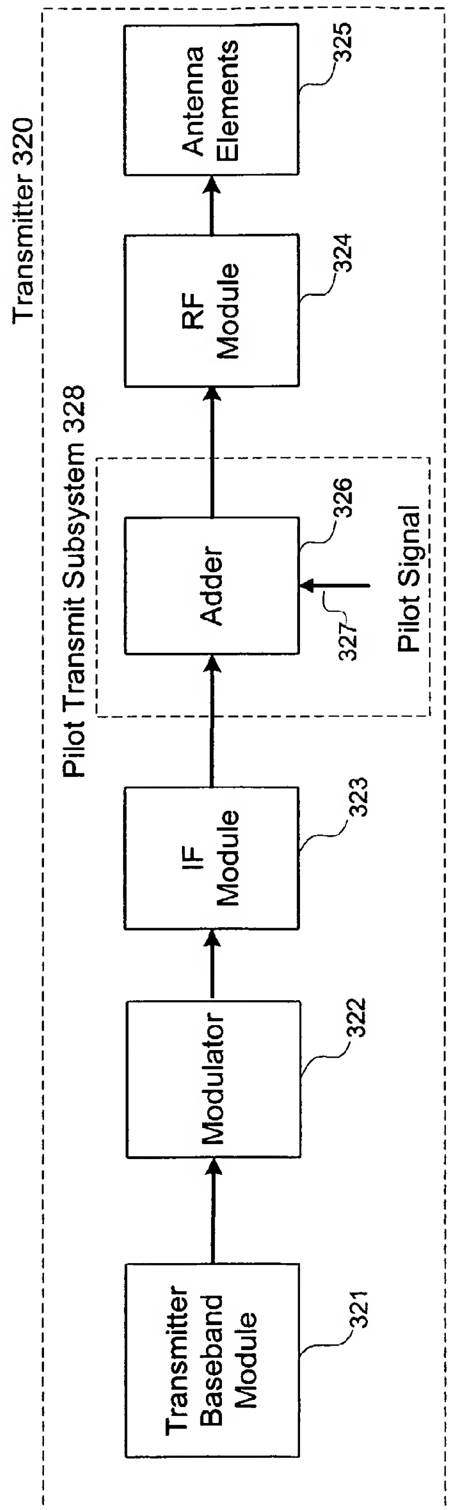
Transmitter 320
Pilot Transmit Subsystem 328
Transmitter Baseband Module
321
Modulator
322
IF Module
323
Adder
326
327
Pilot Signal
RF Module
324
Antenna Elements
325

FIG. 4C

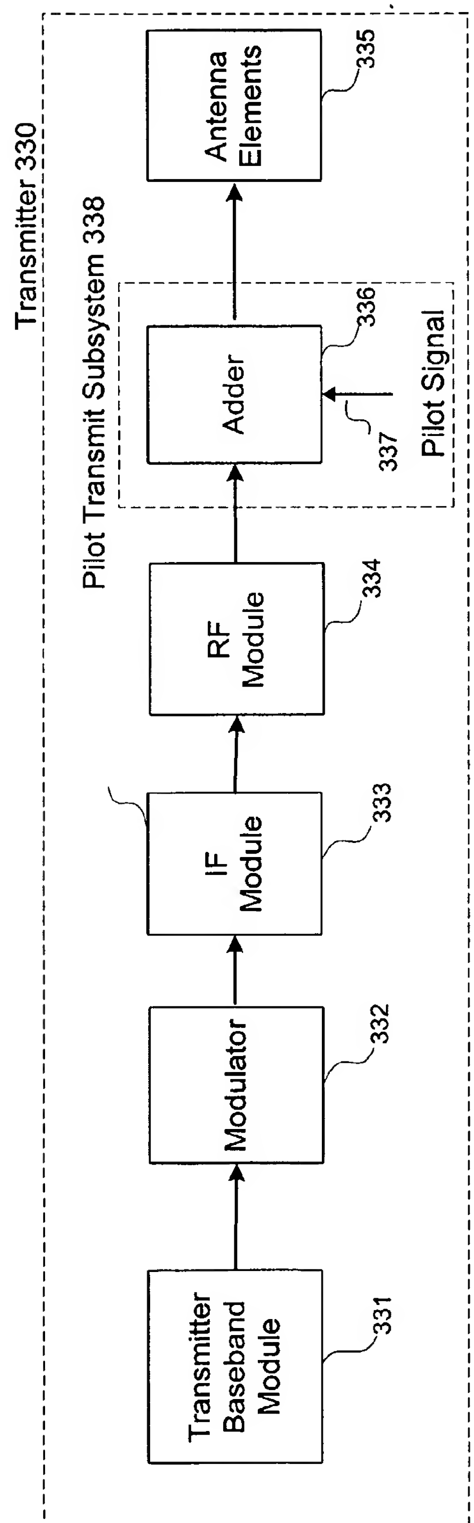
Transmitter 330
Pilot Transmit Subsystem 338
Transmitter Baseband Module
331
Modulator
332
IF Module
333
RF Module
334
Adder
336
337
Pilot Signal
Antenna Elements
335

FIG. 4D

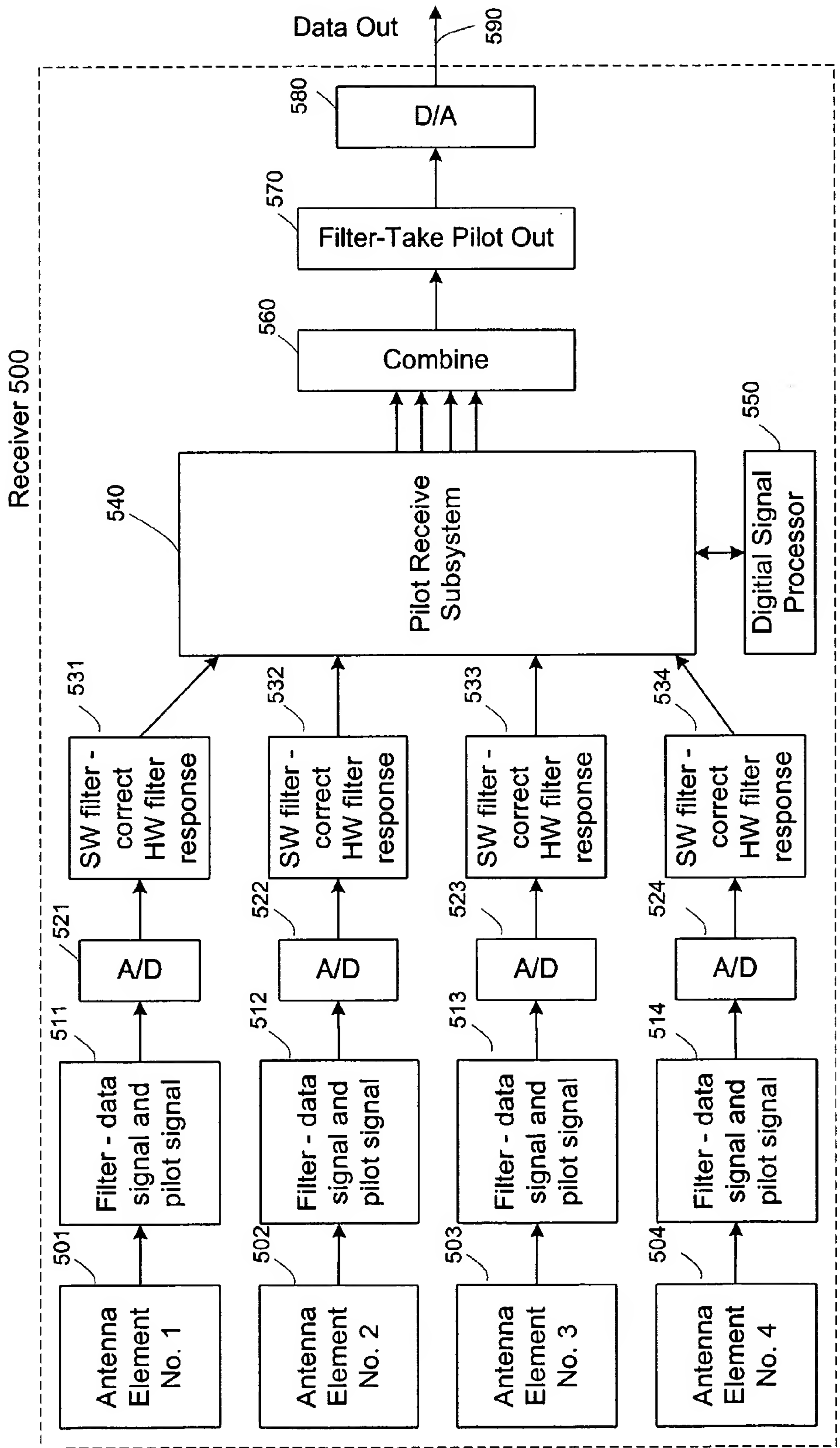
Receiver 500
Data Out
590
580
D/A
570
Filter-Take Pilot Out
560
Combine
540
Pilot Receive Subsystem
550
Digitial Signal Processor
531
532
533
534
SW filter - correct HW filter response
SW filter - correct HW filter response
SW filter - correct HW filter response
SW filter - correct HW filter response
521
522
523
524
A/D
A/D
A/D
A/D
511
512
513
514
Filter - data signal and pilot signal
Filter - data signal and pilot signal
Filter - data signal and pilot signal
Filter - data signal and pilot signal
501
502
503
504
Antenna Element No. 1
Antenna Element No. 2
Antenna Element No. 3
Antenna Element No. 4

FIG. 5

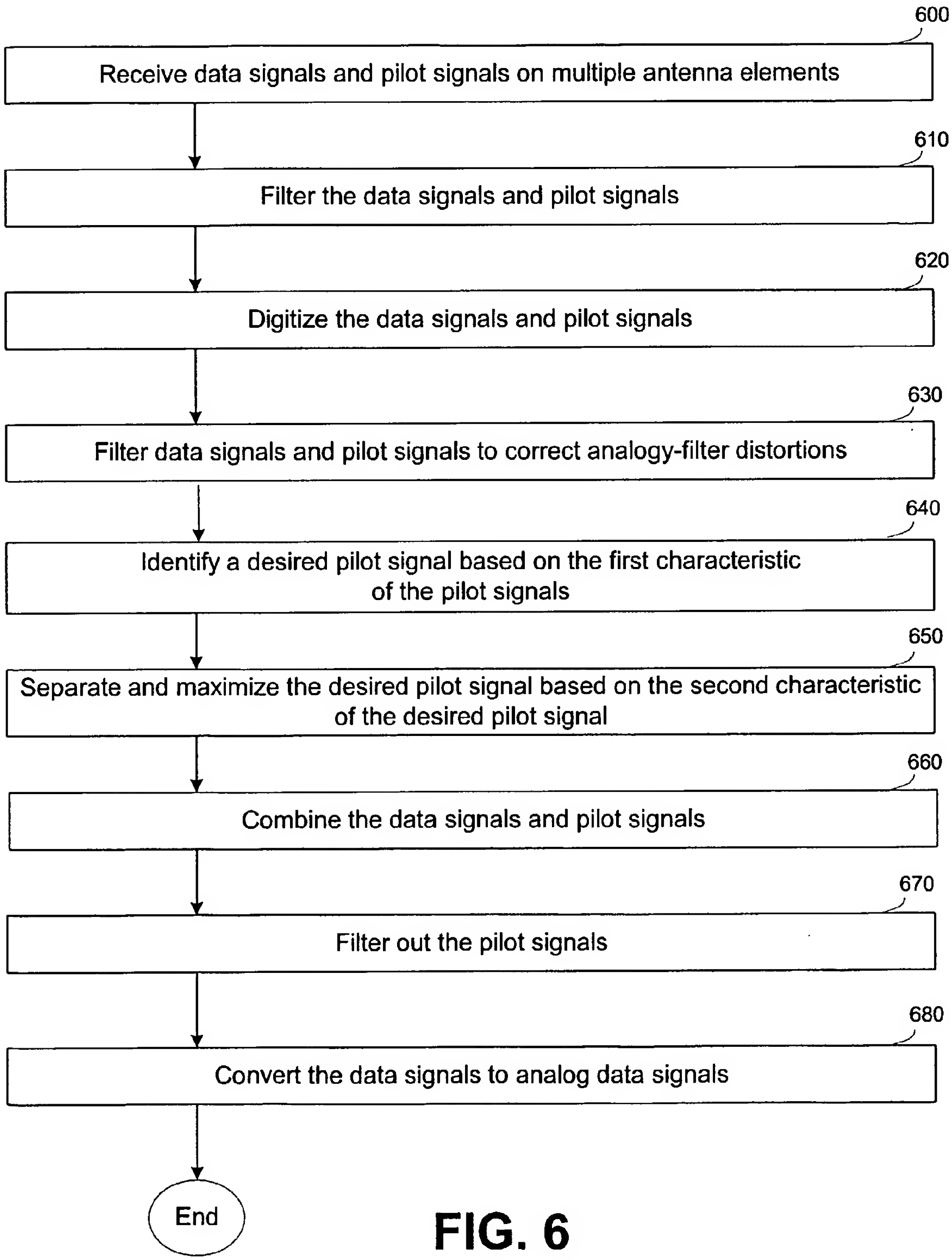
600
Receive data signals and pilot signals on multiple antenna elements
610
Filter the data signals and pilot signals
620
Digitize the data signals and pilot signals
630
Filter data signals and pilot signals to correct analogy-filter distortions
640
Identify a desired pilot signal based on the first characteristic of the pilot signals
650
Separate and maximize the desired pilot signal based on the second characteristic of the desired pilot signal
660
Combine the data signals and pilot signals
670
Filter out the pilot signals
680
Convert the data signals to analog data signals
End

FIG. 6

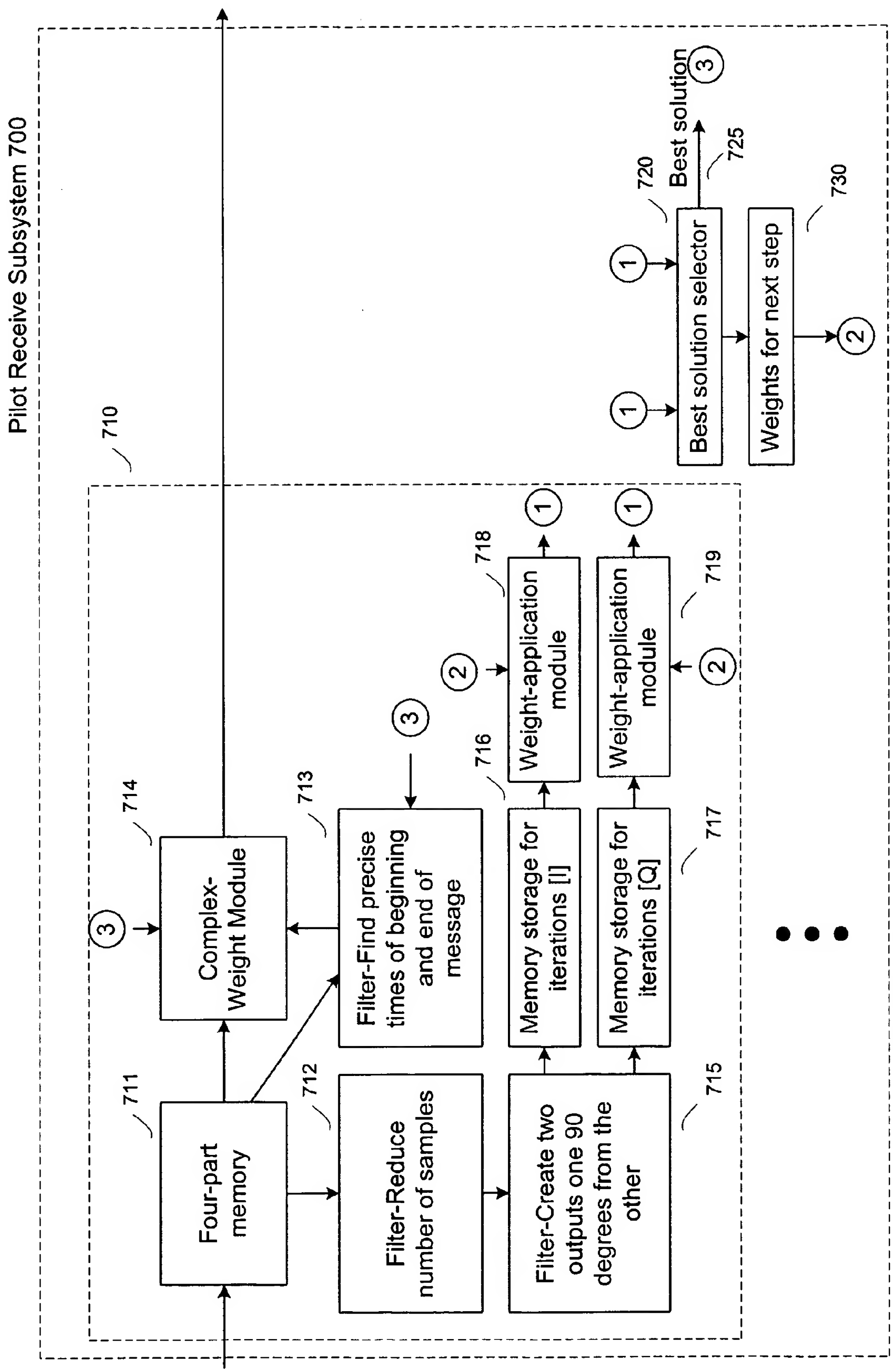
Pilot Receive Subsystem 700
710
711
Four-part memory
712
Filter-Reduce number of samples
715
Filter-Create two outputs one 90 degrees from the other
714
Complex-Weight Module
3
713
Filter-Find precise times of beginning and end of message
3
716
Memory storage for iterations [I]
717
Memory storage for iterations [Q]
718
Weight-application module
719
Weight-application module
2
2
1
1
720
Best solution selector
1
1
725
Best solution
3
730
Weights for next step
2

FIG. 7

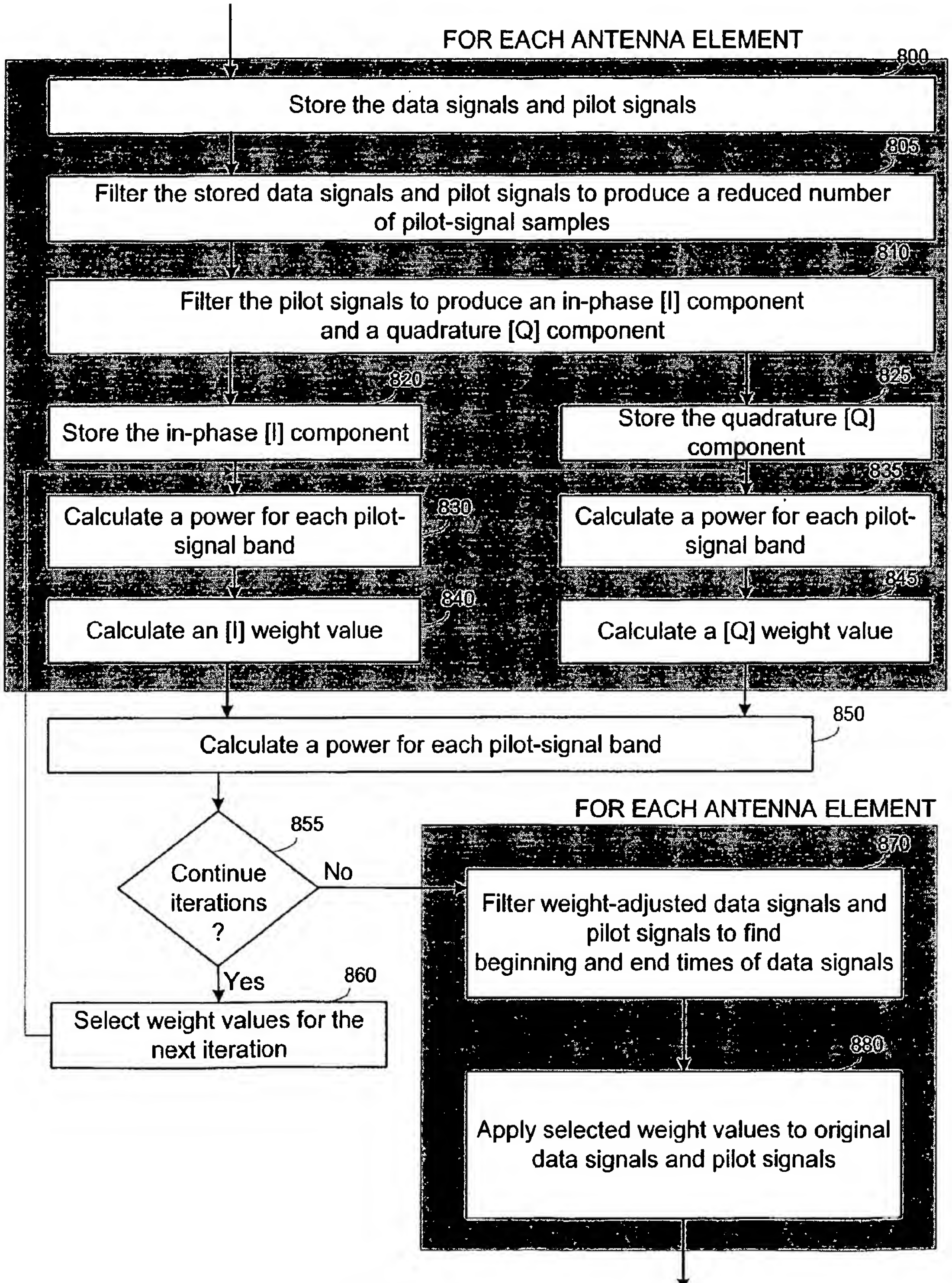
FOR EACH ANTENNA ELEMENT
800
Store the data signals and pilot signals
805
Filter the stored data signals and pilot signals to produce a reduced number of pilot-signal samples
810
Filter the pilot signals to produce an in-phase [I] component and a quadrature [Q] component
820
825
Store the in-phase [I] component
Store the quadrature [Q] component
835
830
Calculate a power for each pilot-signal band
Calculate a power for each pilot-signal band
845
840
Calculate an [I] weight value
Calculate a [Q] weight value
850
Calculate a power for each pilot-signal band
855
Continue iterations ?
No
Yes
860
Select weight values for the next iteration
FOR EACH ANTENNA ELEMENT
870
Filter weight-adjusted data signals and pilot signals to find beginning and end times of data signals
880
Apply selected weight values to original data signals and pilot signals

FIG. 8

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US02/05650

A. CLASSIFICATION OF SUBJECT MATTER

IPC(7) :H04B 7/10; H04L 1/02
US CL : 375/347
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 375/347, 346, 348, 350, 144, 147, 148; 370/ 334; 335, 342,441

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

EAST (pilot, signal optimization, interference (reduction, cancellation), smart antenna)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A,P	US 6,195,342 B1 (ROHANI) 27 FEBRUARY 2001, abstract and col. 2, line 37 to col. 3, line 50.	1-22
A	6,104,983 A (NAKADA) 15 AUGUST 2000, abstract, figure 3, and columns 1-9.	1-22

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
02 JULY 2002	5 AUG 2002

Name and mailing address of the ISA/US	Authorized officer
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